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ham radio magazine

FIRST REPORT: PROPAGATION ON THE NEW BANDS

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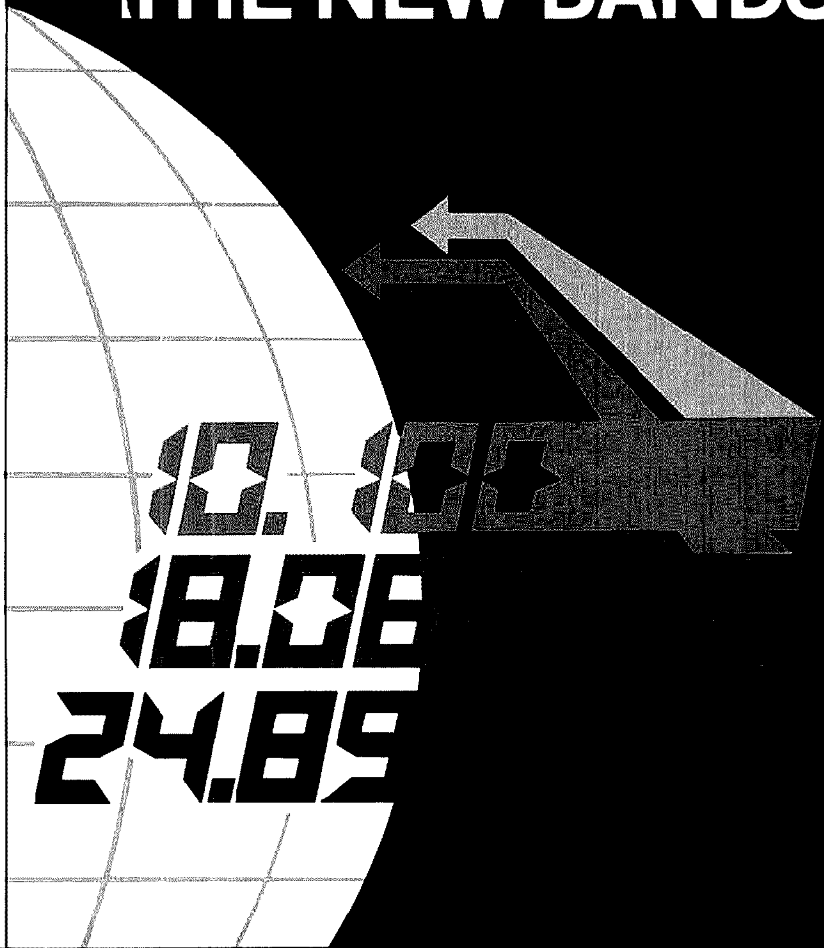
- an interview with Dr. Kenneth Davies of NOAA
- a simplex autopatch
- the logic mate



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contents

REFLECTIONS

An Amateur must have the permission of a TV broadcaster before he can buy equipment to get on a certain Amateur band? That's the decision of a California Superior Court judge in Sacramento County! Of course the issues involved are not nearly so simple as stated above. But whatever the background and however compelling the arguments favoring such a decision, the fact that any commercial enterprise has been given control of the activities of an FCC-licensed Amateur has to be of grave concern to all Amateurs. Here's what happened:

A Multipoint Distribution Service (MDS) in Sacramento, California — California Satellite Systems — went to court for a preliminary injunction against a Sacramento Amateur-equipment distributor and several others for "unfair competition," alleging they were selling equipment for receiving the MDS company's 2150-MHz subscription TV signals, and thus unjustly depriving it of revenue. The judge concurred, despite testimony about the nearby 2300-MHz Amateur band and an in-court demonstration of 2300-MHz Amateur equipment. In granting the injunction, which strictly forbids any of the defendants from selling *any* equipment *capable* of being used to receive 2150-MHz MDS TV, the judge included a proviso that such equipment could be sold to a licensed Amateur or "experimenter in microwave frequency transmissions." The section of the injunction order of serious concern to Amateurs is the requirement that prior approval (at no charge) must be obtained from the MDS company before such a sale can be made.

The case, of course, brings up many questions about individual and business rights. The problems of scrambled UHF TV "pirating," and the private reception of both MDS and satellite TV, are currently receiving plenty of attention from both the trade press and — as in this case — the courts. "Rights" mean different things to different people. There's one school of thought that holds that "anyone who doesn't want me to enjoy his radio/TV transmission should keep them off my property." This viewpoint has considerable popular support. A drive-in theater operator who sued to prevent neighboring home-owners from enjoying his movies from their back yards would be laughed out of court. Yet in St. Paul, Minnesota, a federal district judge ordered a home-owner to take down his MDS receiver or start paying the company a subscription fee!

One remedy for "pirate" subscription-TV reception that has been widely employed is legal sanctions against the manufacture and sale of equipment. In California this led the state legislature to pass a law prohibiting not only the sale of decoders but also of any components that could be used in their manufacture! Court injunctions against decoder sale and manufacture, and even the distribution of decoder kits, have been granted in a number of actions across the country; but in this case the judge has taken the process one onerous step further.

In this case, the application of such legal principles as "prior restraint" and "presumption of guilt" must be questioned. The possessor of a Technician or higher class Amateur license has been granted the right to operate on all VHF and higher frequencies by the FCC. However, in Sacramento this right, at least with respect to the 2300 MHz band, can be subject to the whim of a commercial enterprise — a commercial enterprise that has no interest whatsoever in the furtherance of Amateur Radio but has a very real interest in keeping any and all "amateurs" (note the small "a") and other experimentally inclined individuals away from "their" portion of the spectrum. To their credit, California Satellite Systems' lawyer has contacted Westlink asking the Amateur community for suggestions as to how both their interests and ours can best be protected. Any suggestions can be sent to us here in Greenville, and we'll see they get to the law firm.

However good the intentions of the MDS firm in this case, this decision is one that cannot be allowed to go unchallenged. If it stands, then why not give a channel 2 TV station the right to determine what kind of 6-meter gear is sold in its viewing area, and to whom? After all, 6-meter Amateur interference could drive viewers to other channels, thus depriving the station of revenue. Or why not let a cable system operator establish quiet hours for his area's 2-meter operators, since cable channel E (145 MHz) is used for extra-charge special events on many systems?

The Personal Communications Foundation, the ARRL, and many concerned individual Amateurs are closely following the developments in Sacramento. We must not let such a fundamental intrusion on the rights of licensed Amateurs remain unchallenged.

Joe Schroeder, W9JUV
associate editor

comments

no-code license

In the September, 1982, issue of ham radio, *the publisher and editor-in-chief of* ham radio, *Skip Tenney, W1NLB, and associate editor Joe Schroeder, W9JUV, presented an editorial in which they advocated that Amateurs look into the merits of the FCC's proposed no-code Amateur license. As expected, the response was swift and emotional. By far the majority of those who responded to the editorial were opposed to any kind of code-free license.*

Presented below are excerpts from some of the letters received.

The decision and dilemma of a no-code license is a two part question: fact and emotion. Let's examine some thought-provoking aspects of this question.

What is the value or purpose of Morse code? There is only one! In a marginal reception situation, the ear/thought process can interpret an on/off tone more effectively than the complex communication of the human voice. For this very reason, all military communication specialists are required to learn code. Said simply, anyone using code can get a message through where voice is unintelligible. It is also fact that on/off rf generation equipment (both sending and receiving) is simpler to develop, understand, and build than voice generation techniques.

The second part of the no-code license proposal appears to be the emotional half. I am a ham! Not everyone can be one, because they don't have the determination, intelligence, skill, and guts to work to be a member of our fraternity. I'm pleased to be a member of a group of people who are elite and have done something to make them stand out in a crowd. Our country was founded and has grown on the principle that the best succeed and thank God we can all work to achieve what we want.

Are these questions important? What is the purpose of Amateur Radio? If you are part of a group whose lives depend on "getting the message through" (military or public service), or if you believe in the principle of our ability to "be what we want to be" (freedom), then the answer is yes.

As a final note please observe that I have underlined work many times. True appreciation of anything in life — be it mental, spiritual, or physical — is rewarded directly to the amount of effort expended by the recipient. Anything given to us quickly loses its real value without exception. Time and again we are reminded that each of us only truly enjoys what we have earned. — William E. Martin, WB4KSP.

The code requirement is the only thing keeping more lids from Bashing their way into Amateur Radio. The code represents discipline. Even if a person never uses the code once he gets his license, he has demonstrated the fact that he has the discipline necessary to *be a qualified ham.* Once a person learns the code he does not easily forget it. It is a skill to be proud of having.

I doubt if any ham really believes the code is outmoded, as some computer-oriented proponents would have us believe. I doubt if any ham really believes the code is an impossible stumbling block for youngsters.

There are only two reasons why a person cannot learn the code: 1) lack of interest (he really doesn't want to be a ham); or 2) laziness (he would rather be given something than earn it).

Sorry, gentlemen, you're wrong on this one. If we wind up with a no-code license, it will be due to publishers, manufacturers, and dealers . . . not because of their interest in Amateur Radio, but rather due to their pursuit of the almighty dollar. — Ken Piletic, W9ZMR.

It is perfectly ridiculous to require that computer-oriented young people, and busy young engineers with endless demands on their time, learn the code which they may never use. We urgently need these bright energetic young people in the Amateur service, and we are not attracting enough of them.

Most Amateurs oppose the no-code license because they do not want Amateur Radio to sink into a CB-type chaos. But well-conceived examinations, honestly administered, should safeguard against that. Others oppose a no-code license on the grounds that anyone can learn the code; these well-meaning people simply do not understand the problem.

It is unfortunate that the ARRL, because of the shortsighted view of most of its members, can be counted

continued on page 54

presstop de W9JUV

AMATEUR INTRUSIONS INTO THE OFF-LIMITS 10.109-10.115 MHz portion of the new 30-meter band have resulted in complaints from the governmental agencies affected and concern in the FCC. With Amateurs from other nations using the slots forbidden to U.S. Amateurs, it's easy to slip into violation. When such an intrusion is heard, please warn the violator with a quick long-distance call to help preserve our good relations with the agencies that have legitimate access to those frequencies.

Activity On The New Band Has Been Steady but not heavy, with most users reporting it is more like 40 than 20. DX has been reasonably available, with several U.S. stations reporting more than 40 countries worked and several more near WAS by the end of November.

BURBANK (ILLINOIS) OFFICIALS HAVE FINALLY RESPONDED to their residents' suit challenging the community's anti-tower and RFI ordinance. The city, in a motion to dismiss the complaint, challenged the federal court's jurisdiction and claimed the city has a right to control "nuisances" such as RFI and antennas. A status hearing is set for December 13.

BEACON FREQUENCIES AUTHORIZED BY THE FCC FOR "AUTOMATIC" (unattended) control in its early November report and order are 28.2-28.3, 50.06-50.08, 144.05-144.06, 220.05-220.06, 222.05-222.06, 432.07 - 432.08, and all Amateur frequencies above 450 MHz. Manual control (attended) beacons are permitted on all Amateur frequencies. Beacon transmitters are limited to 100 watts input under the new rules, which become effective January 3, 1983.

ARRL ASSUMPTION OF THE EXAMINATION PROGRAM as proposed in its late October petition is believed to be seeing some modification by the League itself during the comment period, which closed November 29. The League's Executive Committee has also agreed to ask the Commissioners to delay consideration of a no-code license for 18 months, in order to avoid injecting such a controversial issue into the already complex problems that Amateur administration of the exam program is certain to create.

Whether Or Not The League's Exam Proposal becomes the basis of the new exam process, the Commission's own timetable, as well as financial considerations, seem to insure that the transfer will have to be accomplished by next fall.

Amateur Participation In Monitoring And Enforcement activities could well precede our taking over the examination program. ARRL's Interference Task Force has been working on the details of a cooperative effort between FCC's Field Office Bureau and Amateur volunteers for some time, and with minor revisions in the well-established OO program it could be up and running in a few months. Much of such a program would be "advisory" to individual Amateurs, as with present OO notices. However, with respect to deliberate violators, the volunteer monitors' job would be to make it as easy, quick, and inexpensive as possible for the Commission enforcement staff to build an effective case against a flagrant violator.

FCC'S NOVICE EXAM PROPOSAL to permit the examiner to create and grade the written as well as code portions, PR Docket 82-727, has comments due February 15, reply comments March 15.

Comments On The FCC's Logbook Elimination Proposal, PR Docket 82-726, are due by January 14. Reply comments on that item are due February 14.

In A Related Move, The FCC Has Dropped The Requirement that stations in the Aviation Services maintain logbooks.

A NEW RUSSIAN SATELLITE, ISKRA 3, WAS LAUNCHED November 18 from the Salyut 7 spacecraft. At presstime only CW telemetry on its 29.583-MHz beacon had been heard from Iskra 3, which is believed to carry a 21.23-21.27 MHz in, 29.58-29.62 MHz out transponder like its predecessor, Iskra 2. Iskra 3's altitude is about 360 km, and period 91.6 minutes.

Phase IIIB's Launch Is Now Definitely Set For Next April on Arianne's L6 launch. In addition, the Air Force has responded favorably to a proposal that Phase IIIC be given a ride on an Air Force launch some time in 1984. As a result, AMSAT's budget is being stressed severely. Members are strongly urged to renew their memberships promptly. Some worthwhile AMSAT goodies are still available; Box 27, Washington D.C. 20044 has details.

1983 OSCAR Orbital Prediction Calendars that include RS-5 through 8 as well as OSCAR 8 are now available from Project OSCAR, Box 1136, Los Altos, California 94022. U.S. postpaid price is $10; they're $12 overseas.

FINAL WARC IMPLEMENTATION COMMENTS ARE BEING SOUGHT by the Commission in an NPRM adopted at its November 18 agenda meeting. The actual 300-page document and comment due dates should be released by mid December, but those areas of most interest to Amateurs are final rules for 10 MHz, availability of 18, 24, and 902 MHz, and moving the frequency below which CW ability is required down to 30 MHz. No surprises are expected in the Amateur Radio portions of General Docket 80-739; a 60-day comment period is expected.

ELIMINATION OF CB LICENSING IS EXPECTED TO BE DECIDED ON by the commissioners in an early December meeting. Expectations are, that despite strong opposition from CB user groups such as REACT, requirements for both CB and radio control licensing will be dropped.

NEW THIRD PARTY TRAFFIC AGREEMENTS WITH ST. VINCENT (J8) AND GRANADA (J3) have been announced by the FCC, just in time for Caribbean winter holiday travel.

first report on new band propagation

Experiments and observations on the 10, 18, and 24 MHz bands

The Federal Communications Commission licensed the experimental station KK2XJM on August 25, 1981, for the purpose of conducting experiments on the 10.100-10.150, 18.068-18.168, and 24.890-24.990 MHz WARC bands. These bands are intended for Amateur use in the future. A license on these bands authorizes operation at a maximum of 30 watts ERP and 2.8 A3J emissions. [U.S. Radio Amateurs were permitted to use 10.100-10.109 MHz and 10.115-10.150 MHz, at a maximum power level of 250 watts with A1 or F1 transmissions, as of October 28, 1982, at 3 p.m. Eastern Daylight-Saving Time.]

One main reason for the experiment on these bands was to gather and analyze reports received from (mostly Amateur) listening stations. The analysis techniques we used, and the results obtained from the reports, are presented in this article.

Because the results of the experiment depended greatly on listener response, we generated and distributed publicity announcing the experiment and its authorization by the FCC, and we described the information needed from listeners and how it should be reported. This publicity was followed at intervals by detailed schedules of our broadcasts. *QST* and *World Radio* published the complete announcement, *ham radio* covered the information in an editorial, and extracts and schedules were also published in *73*, *CQ*, *Florida Skip* (a regional magazine), and several local club newsletters.

The response we received from listening stations was encouraging. Many people phoned or sent letters, requested schedules, asked for information on the equipment to be used, and wanted to know how to provide reception reports.

preliminaries

Before starting the experiment, we conducted a number of sweeps on the 10, 18, and 24.5 MHz bands to determine signal levels and types and to select operating frequencies. **Fig. 1** shows a result typical of these scans. Signals are shown as blocks for teleprinter, multi-channel signals, and voice or music, and as lines for unmodulated carriers. Some of the carriers represent no-information transmission periods of stations that were active at other times. We found gaps at times, but all frequencies on 10 MHz were in use at some period of the day.

Conditions were similar on 18 and 24.5 MHz, but the average number of stations on the air was lower, and periods of heavy use were shorter. It was not uncommon to find only six signals on these bands, half with no modulation.

A better picture of band use developed as the experiment progressed. Spectrum scans were made by W1RN at the ARRL Lab in Newington, Connecticut, as an aid. A pattern of full occupancy during others was evident; we also noted that a few signals were very strong, but there were long periods when most signals were not much above the noise level.

We couldn't identify all the signals present; stations appeared without identification, transmitted coded signals by voice or CW, and then disappeared. We didn't try to copy teleprinter signals, which alternated as keyed or non-keyed carriers without Morse transmissions.

After studying the scan results, we scheduled initial operations on 10.140, 18.108, and 24.930 MHz. These are each 40 kHz above the low edge of the WARC bands. This uniformity was actually a low priority; the prime consideration was to avoid frequencies occupied either by very strong signals or by any signals most of the time. And we wanted to select frequencies most likely to be detected by typical Amateur installations, as well as those on which interference would be a likely problem. Determining the interference potential to existing band users by an Amateur signal was one of the main reasons for conducting this experiment.

By R.P. Haviland, KK2XJM (W4MB), 2100 S. Nova Road, Box 45, Daytona Beach, Florida 32019

Although it soon became clear that there was no real reason to use frequencies other than those first selected, it did appear that some attempt should be made to simulate the Amateur condition of no assigned frequency. We did this by introducing a small offset from the selected frequency. Normally the offset was kept to a fraction of a kHz, but we did use a few offsets of up to 5 kHz.

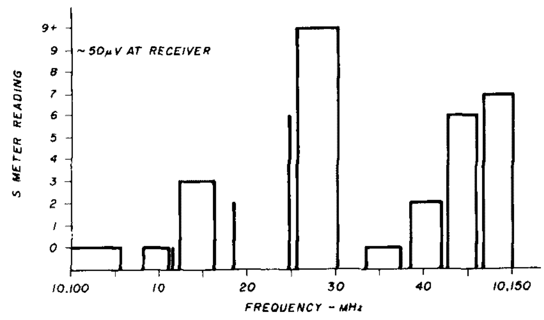


fig. 1. 10-MHz-band signals on September 20, 1981. Wide blocks are single or multi-channel Teletype™, the narrow block is SSB, and the lines represent carriers.

experimental equipment configuration

The equipment we used is summarized in **fig. 2**. The transmitter is an SBE 33, using new band crystals, and rf coils paralleled by additional inductors to provide the necessary tuning range. The tune circuit was modified to provide an adjustable carrier level and to permit switching or keying.

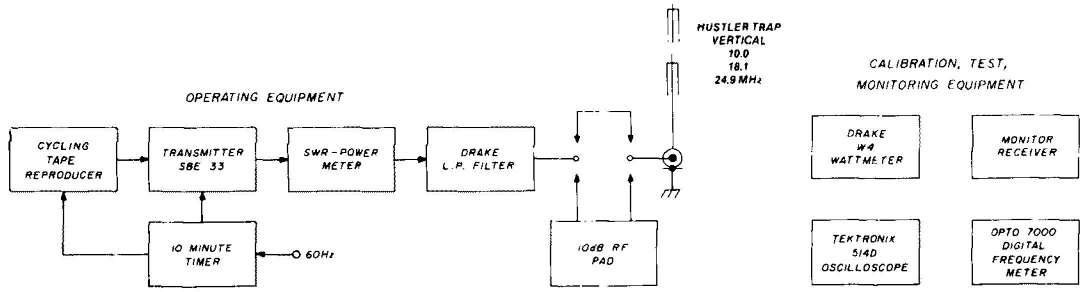


fig. 2. Block diagram of experimental configuration.

We retuned the antenna, a commercial (Hustler) trap vertical, to the center of the WARC bands by adjusting the capacity sleeve of each trap. Section lengths were adjusted for minimum SWR at the selected frequencies. The antenna was mounted at the center of a 36 × 56 foot metal roof. The ground connection to the roof was primarily capacitive: a screen approximately three feet square with six end-terminated radials eight to thirty feet long. An rf check of the roof showed essentially uniform current distribution at the perimeter.

The transmitter operation was controlled by a ten-minute timer, the minutes indicator of an experimental digital clock. We used a nine-minute carrier followed by one minute of identification by USB voice. The identification was made by a recorded voice announcement of fifteen or thirty seconds' duration on an endless tape. For ease of copy, the XYL (W4LDY) made the tapes. The identification tape was not locked to time.

We monitored power and frequency continuously. A 10-dB pad (2-watt carbon resistors in oil), was inserted in the transmitter output for low power. This also served as a dummy antenna.

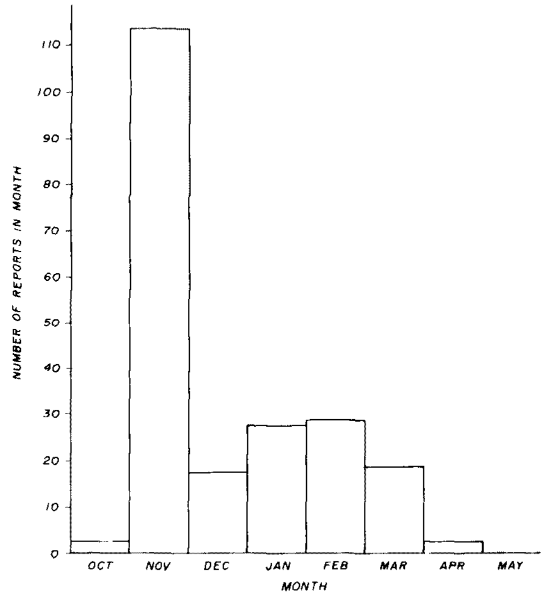


fig. 3. Distribution of reports by month.

Band use and emitted signal quality were monitored by a transceiver, originally a Kenwood TS-820, later a TS-830. We began operating on October 3, 1981, at a power level of 3 watts ERP on 10.140 MHz. The signal was a single steady carrier interrupted every ten minutes, beginning at two minutes past the hour, by a voice announcement. We operated continuously from October 4 through October 12, and then for seventy-two hour stretches (Friday through Sunday, UCT) beginning October 19 and continuing till September, 1982.

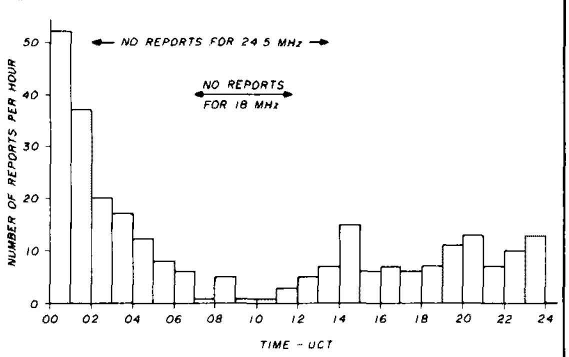


fig. 4. Distribution of reports by time of day. The period from 02 to 12 hours is nighttime at transmitter site.

We increased power to 30 watts on November 3, and instituted a regular rotating schedule for operations on all bands on December 2. During our first four weeks on the new bands, operation was also on 3 watts ERP, rising to 30 watts ERP thereafter.

quantitative results — general

A total of 168 individual reports were received. Of these 131 were for a single time and frequency. The remaining thirty-seven were for multiple times, up to approximately thirty. The total data base consists of 274 entry items: this number is smaller than we'd hoped for, but good enough to give reasonably accurate results.

The distribution of reports over time is shown in **fig. 3**. The small number received in October is a result of low power and a lack of publicity. Changing these factors resulted in a great increase in the number of reports for November. Thereafter, we believe the novelty of reporting was gone, and only the truly dedicated have continued to report. All reports are acknowledged with a special QSL card.

quantitative results — effect of times

The distribution of reports correlated with time of

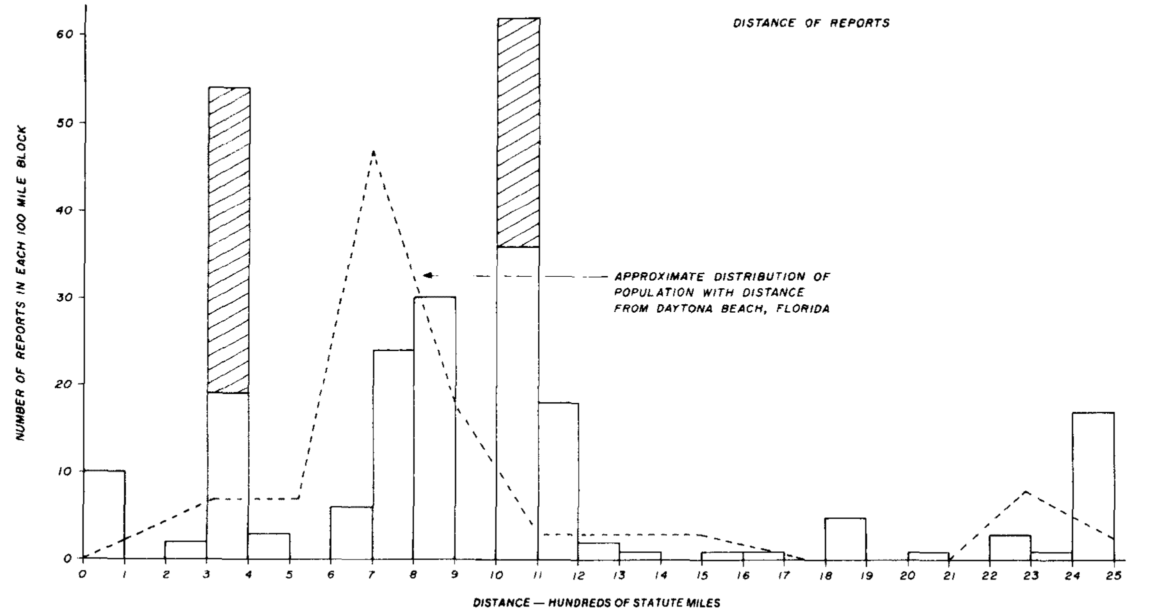


fig. 5. Distribution of reports by distance from the transmitter site. The shaded areas represent reports by two dedicated Amateurs. The dotted curve is the distribution of U.S.A. population from the transmitter site. Note the strong suggestion of skip zones.

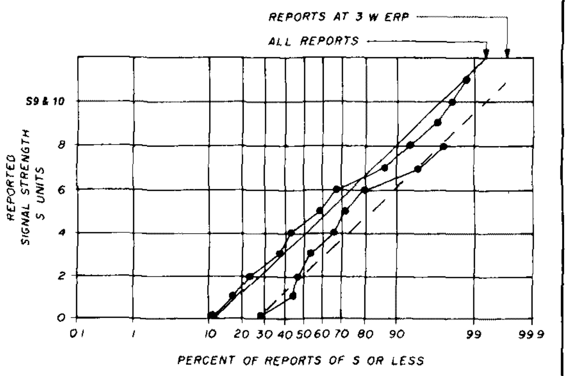


fig. 6. Distribution of signal strength reports. The horizontal scale is for the normal or Gaussian distribution.

day is shown in **fig. 4**, by block for each hour. The reason for the large number of reports at 00-01 hours UCT appears to be that those hours are good for listening (early evening for the U.S.A.) – and also improved propagation at 10 MHz. The low between 07 and 11 hours UCT is understandable, since most listeners are sleeping, and propagation during those hours on 18 and 24 MHz is not as good, as indicated by the no-report bars. The number of reports on these bands, however, is not a major factor in the overall distribution.

The curve indicates that the 10-MHz band will be usable almost twenty-four hours a day during part of the solar cycle. The band at 24.5 MHz resembles 10 meters in performance, being dependent on daylight and high solar activity.

quantitative results – distance effects

The number of reports as a function of distance is shown in **fig. 5**; the reports are separated into zones one-hundred miles wide. (Note that in two of these zones, from one-third to two-thirds of the report entries are from a single Amateur station. The influence of this repeated reporting can be estimated by disregarding the cross-hatched areas of the bars at 300-400 and 100-1100 miles.)

The data strongly suggests the presence of skip zones, with maximum signal strength at 350, 950, 1850, and 2450 miles. This finding, however, could be the result of uneven population distributions. To check this, we calculated the distance from the transmitter to each state capital (we assumed that the largest population concentration of the population of the state is at this distance). The population distribution that results is shown as the dashed curve in **fig.** 5. (While a distribution of Amateurs with distance would be more representative, the percent of Amateurs and shortwave listeners does not appear to vary widely from state to state.)

Comparing the population and report distributions confirms that skip processes are at work here. The block at 350 miles appears to correspond to one-hop (sporadic) E, those at 950 and 1850 to one and two-hop E, and the block at 2450 miles possibly to three-hop E, but more likely to one-hop F-layer propagation. These reports are for all the bands, but they are dominated by 10-MHz reports. The number of reports at 18 and 24.5 MHz were too few to permit the evaluation of the effects of distance.

signal strengths

The distribution of signal strength reports is summarized in **fig. 6**. For all reports on all bands, the median signal report is approximately S4, and nearly fifty percent of the reports fall between S2 and S6. This distribution approximates the normal, or Gaussian curve, very closely, although in principle the curve is bound at the low signal end and essentially unlimited at the upper.

Reports at the three-watt input level are also plotted, and lie almost exactly two S-units, or a nominal 10 dB, below the curve for each. The number of reports at this power level is not, however, sufficient for a really good estimate to be made.

The signal strength distribution over the three bands is shown in **fig. 7**. The median signal on the two higher bands lies approximately one S-unit below signals at 10 MHz, and shows somewhat greater propagation variability. Under strong signal condi-

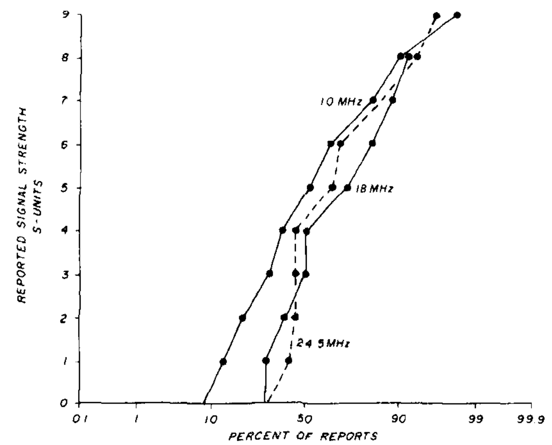


fig. 7. Comparison of distribution of reported signal strengths for the three bands used in this experiment.

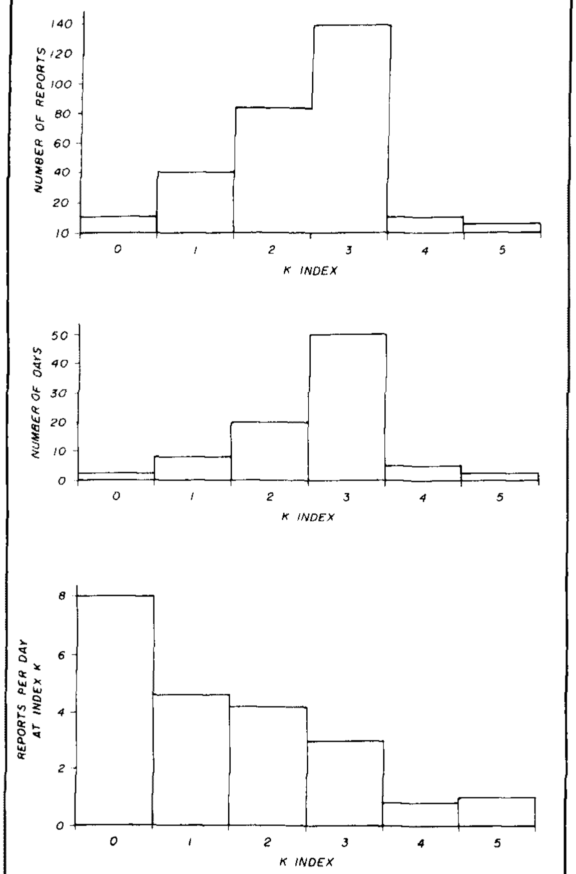


fig. 8. Effect of propagation disturbance on the number of reports received. The bottom curve, giving the reports per day at a given index, is the best indicator of propagation effect. Curves for 18 and 24.5 MHz are similar.

tions, however, levels on all the bands are essentially the same.

These observations suggest that signals on 10 MHz are less affected by most propagation disturbances than will those on the higher bands. (This is also noticeable on 14 MHz, when compared with 21 and 28 MHz.) Considerably more data is needed, however, to confirm this factor. Listening stations' use of less efficient antennas (than a resonant half-wave) might account for the lower signal levels reported.

solar and geomagnetic activity

The National Bureau of Standards' K index was used to measure propagation conditions. Although there were a few periods of very good or very poor propagation, the general conditions ranged from average to below average.

The effect of propagation conditions at 10 MHz is shown in **fig. 8**. The largest number of reports received were for $K = 3$, with a very few reporting K exceeding 3. The ratio, or number of reports per day of a given index, appears to be a better measure of the effects of propagation. The bottom chart shows this, and indicates a marked decrease in the number of reports received as the K index increases. In comparison with the excellent conditions $(K = 0)$, average conditions $(K = 2)$ reduce the number of reports to one-half, and poor conditions $(K = 4\text{-}5)$ reduce the number to around one-eighth.

The pattern for the 18 and 24 MHz bands is similar, with a greater reduction in reports as propagation becomes poorer. This reduction corresponds to subjective observations on the effects of poor conditions.

The effects of conditions on signal strength are shown in **fig. 9**. Three conditions are adduced: excellent, average, and poor, which correspond to K index values of zero, 1-3, and 4 or more, respectively. Signal level differences of 1 to 1.5 S-units between groups are indicated. (Separate observations suggest that F-layer propagation is very poor at $K = 4$ and E-layer at $K = 5$ or greater. Many more observations are required to separate out the various propagation effects.)

receiving equipment

In general, the receiving equipment used by reporting stations was not as sophisticated as the average Amateur band receiver. Most receivers were either all-band receivers, often of the WW II class, or recent-model receivers with WARC band provisions. However, that is not considered to have been a major factor in signal reception.

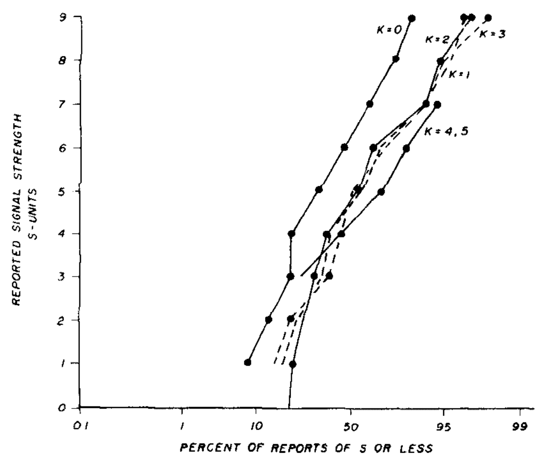


fig. 9. Distribution of reported signal strength by propagation index.

Antennas used by reporting stations (not always optimized for the new bands) included:

inverted-V	76
multiband dipoles	49
long wire	31
verticals	19
wire, (not long)	12
Yagi beams	11
quad-delta beam loops	4
indoor antennas	4

Of these, the long wire and multiband dipoles, and possibly the inverted vees, can show gain, but the rest will give near isotropic performance or worse. For example, the long wire antenna reports were about one S-unit stronger, on the average, with half the reports between S3 and S7. The other half of all reports were between S 1.9 and 6.1, indicating a 5 dB gain for the wire antennas. This suggests the average antenna used in reception was not far from zero dB gain. Indoor antennas could easily be twenty or more dB below isotropic.

Studies reported in *Ham Radio HORIZONS*, July, 1980, page 18, indicate that few Amateur stations on the high-frequency bands use antennas of essentially dipole gain. The average station uses a beam of about 7-8 dB gain, a three-element Yagi or equivalent. Combined beam plus height may run to 15 dB or so for exceptionally equipped stations.

These factors indicate that signal reports in a well-established Amateur Service on these bands will be higher than reported here, resulting from improved equipment alone. The antennas likely to be used would be loaded two or three-element Yagis, showing an effective gain of about 5 dB. Accordingly, it appears that our results should be adjusted upward by about 5-7 dB when applying them to typical (future) operational conditions.

interference

Evaluating interference is an important part of this experimental program. First, interference caused by the experimental station to existing users: No reports of interference to an existing band user were received. This is not surprising; it was the predicted result. In fact, the power limit of 30 watts ERP was set to give a 10-dB margin over the "negligible risk of interference" point of typical existing services, based on CCIR values of required S/N, error rates, fade margin, and typical installation. That no interference has been reported is a good indication that calculated interference values are at least reasonably close to experienced values.

It should be noted that operating frequencies were chosen to be in a low occupancy, but not clear, part of the band, just because of the need for interference data. That this was an effective choice is shown by

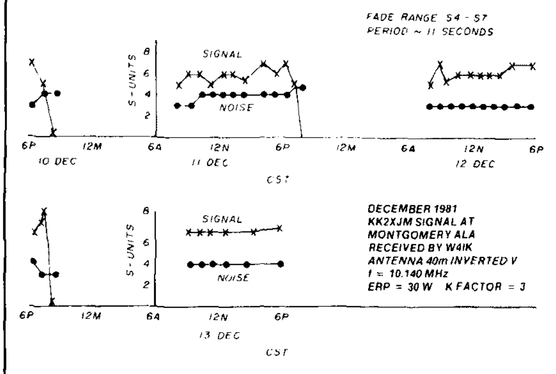


fig. 10. ***Signal and noise levels at a station 350 miles from the transmitter. Note the sharp decrease in signal shortly after sunset.***

the fact that over twenty percent of the reports indicate interference to the experimental signals by existing users.

The relationship between signal, noise, and interference is summarized in **fig. 10**. It shows typical signal and noise levels for a 10-MHz signal received via one-hop E transmission. A decrease in signal strength at sunset is notable. Even at this short distance, 380 miles, the maximum S/N ratio was no more than 30 dB, the usual value being 10-15 dB. Signal levels at this distance should be regarded as marginal for SSB, good for CW.

The interference potential is illustrated in **fig. 11**, which shows a reported band scan for a period of no beacon operation, start-up, and sunset fade. At no time did the maximum beacon signal exceed the level of adjacent signals, being 2-3 dB lower under best beacon-signal conditions. The sunset fade did change the levels of adjacent stations somewhat, when compared with earlier conditions, but the effect was very small compared with that on the beacon signal. (The other signals present were not identified. Since the strength did not usually change appreciably at sunset, they might have arrived via the F layer.)

other experiments

The experimental program described here is one of several in progress. Other stations involved are these:

KK2XGH	Silver Springs, Maryland
KM2XDU	Rye, New York
KM2XDW	Menlo Park, California
KM2XDX	New York, New York

Experimental transmissions are continuing. Major changes that have been made at KK2XJM are power increase to 100 and 300 watts ERP, and scheduling

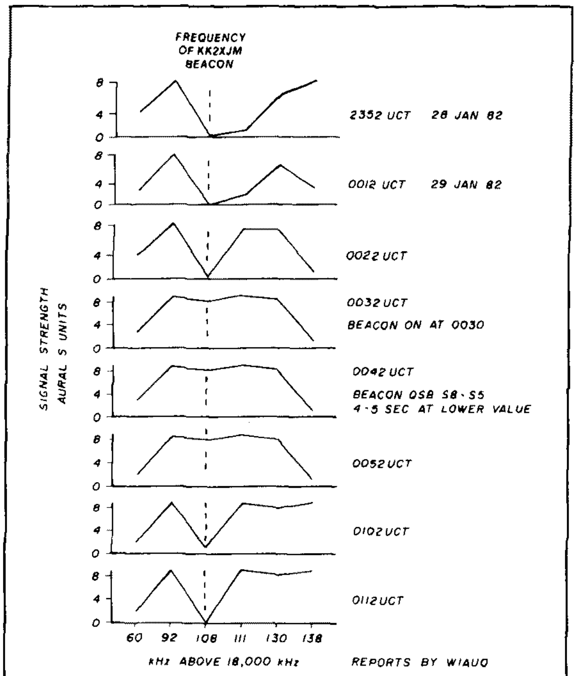


fig. 11. Behavior of experimental signal, and adjacent signals at a distance of about 1200 miles.

band use to approximate the OWF to selected areas. Test results have not yet been compiled for these changes.

conclusions

Our findings are summarized by noting that no unexpected conditions occurred. Propagation effects and signal levels are mostly very close to the predicted average values. One possible conclusion is that a viable and reliable international (greater than 2500 miles) Disaster-Emergency Communication Service* on 10 MHz cannot be established using SSB at 30 watts ERP. This is clearly shown by the paucity of international reports, by the low S/N ratio at 1100 miles and low signal level at 2500 miles under average conditions, and by the lack of service reliability.

This finding confirms the propagation calculations we filed with the experimental license application. These show, for just usable service, operator to operator, average conditions, with a dipole receiving antenna, the following transmitter ERP requirement:

path	time	mode	power, CW	power, SSB
London-Azores	Sunset	E-layer	0.4 W	26.3 W
Washington-Paris	Noon	F-layer	8.2 W	213.8 W

*The benefits of such a service was a major factor in securing these new bands as an Amateur allocation.

For a reliable service, power should be increased by about 10 dB for quality, by a further 10 dB for fade margin, and by another 6 dB for worldwide paths.

It appears that power levels of 1500 watts ERP are needed for reliable worldwide CW disaster/emergency service with dipole antennas (typical of emergency conditions). The level could be reduced to about 300 watts if inter-station relay or a beam antenna for *reception were used*. A highly reliable worldwide SSB emergency service appears to require unreasonably high powers, although a useful regional service could be established with powers of 200-600 watts ERP.

A second conclusion is that a power level of 30 watts ERP, either AØ or A3J, is below the level which would cause harmful interference on the 10 MHz band. This finding confirms interference calculations. For example, consider a path with 1000 km E-layer winter night propagation, a total of ten interfering signals per 0.3 kHz (CW) or 2.8 kHz (SSB), and CCIR recommended performance. For a signal service reliabilty of 99.99 *percent (FSK) or ninety percent* (commercial voice), the allowable ERP of *each* interfering station is:

situation wanted/unwanted	CW/CW	CW/SSB	SSB/CW	SSB/SSB
allowable ERP	154 W	912 W	309 W	1830 W

The allowable radiated interference level varies roughly as the square root of the number of stations of equal power (or as the RMS equivalent if different power), due to statistical variations, in modulation, timing, arrival phase and path attenuation. Interfering signal power could be increased about 20 dB at noon. A 10-dB increase in the allowed interference level would reduce service reliability of the desired signal to about fifty percent.

The fact that SSB emissions may be allowed higher power has surprised some. This is due in part to the difference in duty cycle: − 17 dB assumed for the normal voice versus − 3 dB for average CW. Another difference is in the effect of the number of signals present in the passband, and the percentage of the total radiated power accepted by the receiver. In a service with distributed interference sources and random frequency use, the measure of interference potential is power per unit of bandwidth. Emitted bandwidth alone is not important.

Several weeks of operation at 100 watts and 300 watts ERP have passed since the period this report was based on, again with no reports of harmful interference. This offers further confirmation of the above evaluation. Additional results will be reported on later.

ham radio

a fixed-tuned LF converter

An add-on project to let your 80-meter receiver explore new worlds

Many commercial and military signals can be received by using this simple converter. It is compatible with most commercial transceivers where a direct frequency readout of the 80-meter band is displayed on the tuning dial. A block diagram of the system is shown in **fig. 1**. For digital-readout transceivers, read the frequency by subtracting 3.5 MHz. The converter covers as low as 3 kHz and up to 240 kHz, the low-end cut-off frequency being determined by the receiver used as a tunable i-f (see **fig. 2**).[1]

the circuit

Parts for the converter were available in my junk box and through local distributors such as Radio Shack. The converter is designed to be operated with a long wire antenna, 100 feet or greater, although results have been fairly good with shorter wires (with some sacrifice in weak-signal reception). An active antenna can be used to increase sensitivity.

As shown in the circuit diagram, **fig.** 3, the input circuit is a six-element Chebyshev lowpass filter which is directly coupled to the FET mixer, Q_1. The CR1 and CR2 diode arrangement protects Q_1 from strong transients such as those created by lightning. The local oscillator is a simple 3.5-MHz Pierce crystal oscillator. An output match to the receiver is provided by an emitter follower, Q_2. Total current drain is less than 3 mA. A 9-volt transistor battery can provide over one year of service under normal use. No degradation of operation was observed even when the battery output dropped to 4.5 volts.

construction

The complete converter, including a 9-volt transistor battery, is mounted in a standard 5 × 2½ × 1½-inch (130 × 65 × 39 mm) utility box. A BNC connector can be used for the i-f output (I used a readily available SMA connector), and a standard banana jack serves for the antenna input. Circuit layout is not critical — a 2 × 3-inch (51 × 76 mm) perfboard can be used for component mounting.

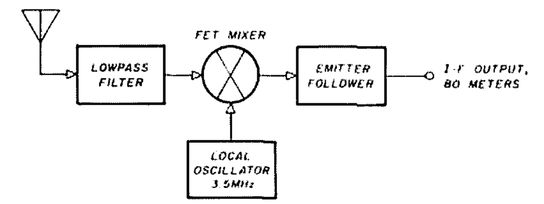


fig. 1. Block diagram of the lf converter.

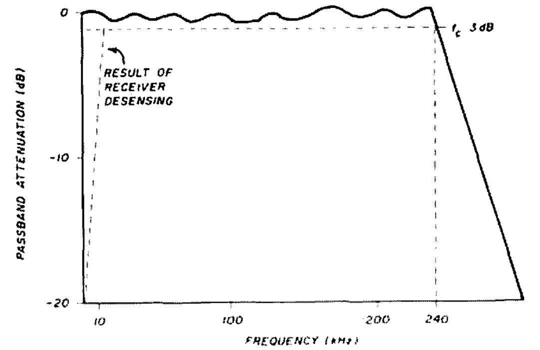
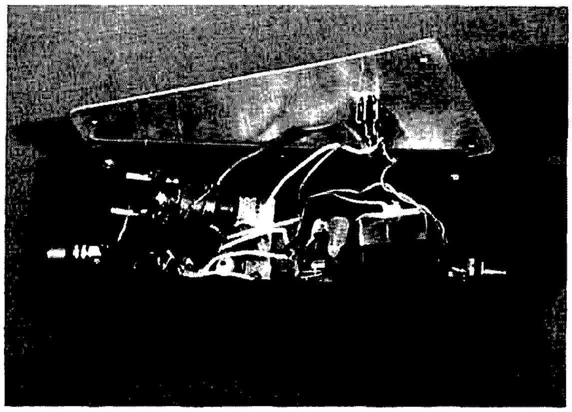


fig. 2. Front-end filter response and tuning range of the 80-meter receiver.



The perf-board and components easily fit into a small utility box with plenty of room for a 9-volt battery.

By S.J. DeFrancesco, K1RGO, 17 Jeffrey Rd., East Haven, CT 06512

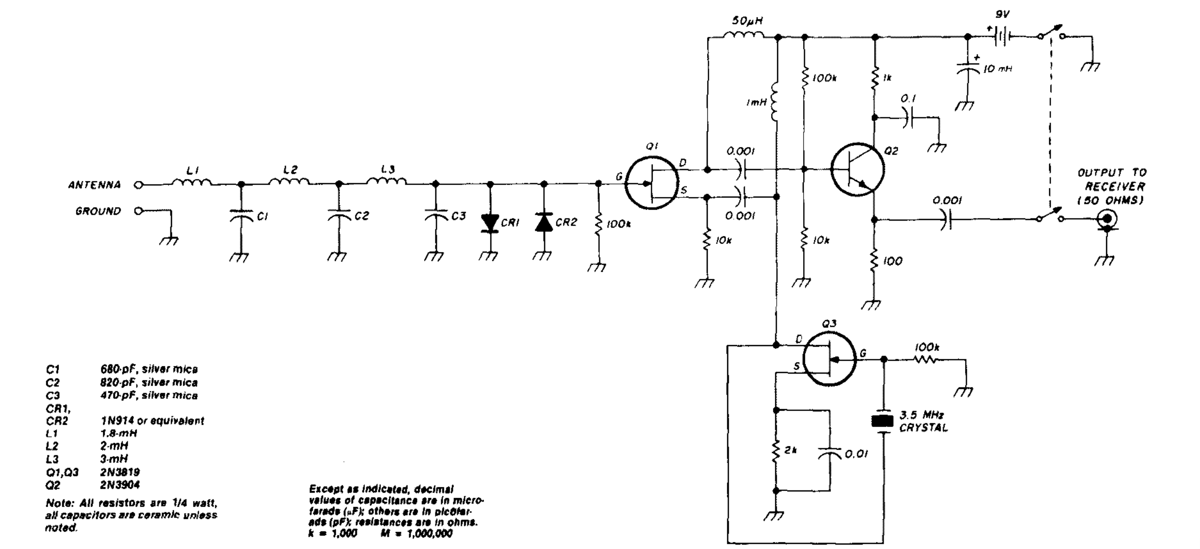


fig. 3. Schematic diagram of the lf converter. Coils used were from surplus two-way radio i-f cans (292 kHz). Rf chokes of proper value will work as well.

table 1. Some signals that have been heard from the East Coast in the LF Spectrum.*

Frequency	Station	Location
12 kHz	OMEGA	–
16-17 kHz	JXZ, GBY	Norway
60 kHz	WWV time signal	Colorado, USA
100 kHz	Loran C	–
120-150 kHz	TTY, CW, Military	East Coast, USA
150-180 kHz	Foreign broadcasts	Mostly European
160-190 kHz	1 watt experimental beacons	East Coast, USA (beacons received over 200 miles)
192 kHz	SFI	San Francisco, CA
194 kHz	TUK	Block Island, RI
215 kHz	CLD	–
233 kHz	GN	–
240 kHz	MMK	–

***Editor's note:** For those interested in studying low-frequency propagation, a good source of frequency and station-location information is the Sectional Aeronautical Chart used by aircraft pilots. Some of the non-directional beacon stations are on frequencies as low as 200 kHz, and they operate continuously. All have an identifier (usually on CW), and some provide *local wind, weather, or air-traffic information by voice announcements. Go* to your nearest airport and ask the FBO (Fixed-Base Operator – he sells/rents/services small planes) for the VFR chart that covers your area. Get one with a scale of 1:500,000. Each chart has a guide to adjoining areas, so you can pick up enough to cover a large part of the country for a small fee. If you are not familiar with the terms used on the chart, ask the FBO to point out a couple NDBs for you. They are usually indicated by a dot, surrounded by a circular dotted area, and have the location/frequency/identification legend enclosed in a nearby rectangular magenta-colored box.

evaluation

I have obtained very good results using the Yaesu FT101EE receiver and my 160-meter antenna (260-foot long wire). The antenna is connected directly to the input of the converter, and the auxiliary antenna input on the Yaesu is connected to the converter's i-f output. Light-dimmer noise is greatly reduced by the noise blanker in the Yaesu, making perceptible most of the weak signals that were otherwise masked by the noise.

In the 0 to 12-kHz portion of the spectrum, which was at one time used for missile detection, you can hear 60-Hz power-line harmonics, as well as whistlers, chirps, and pops caused by atmospheric phenomena. Some of the activity, starting from 12 kHz (3.512 MHz on your receiver) to 240 kHz, is shown in **table 1**. Low power (1-watt CW) experimental beacons and QSOs can also be heard in the 160-190 kHz band. I have heard HS, Fairfield, CT, 165 kHz; FY, Long Island, NY, 186 kHz; KEN, central NJ, 185.6 kHz; BIL, central NJ, 185.9 kHz; and WI in Owings, MD, 186.2 kHz. All signals were very readable (depending on local noise conditions, of course).

reference

1. *Handbook of Electronics Calculations*, Kaufman and Seidman, Filter Design Chapter.

ham radio

technical forum

Welcome to the *ham radio* Technical Forum. The purpose of this new feature is to help you, the reader, find answers to your questions, and to give you a chance to answer the questions of your fellow Radio Amateurs. As a new feature, the Technical Forum will be shaped by the type and number of letters we receive from you. Do you have a question? Send it in!

I am interested in developing an acoustical filter for CW. In the 1920s or early '30s, in QST, there was an article about an acoustical filter, but I have been unable to find that article again.

What I would like to make is an acoustical filter that I could add to ear plugs. The early QST article was just such a device. Since my ears are not the same with regard to loss, I could make a pair of filters and match the gain to each ear.

I realize that my problem is of a specialized nature but the subject of acoustical filters is still an interesting answer to CW filtering. — Howard O. Lorenzen, W7BI.

Your memory is excellent. The article you are referring to appeared in the August, 1928, issue of *QST* on pages 23-29. It featured a detailed explanation of the theory behind and the operation and application of these filters. It is entitled "Acoustic Wave Filters and Audio Frequency Selectivity," by R.B. Bourne, 1ANA.

The article clearly illustrates the wealth of technical information developed fifty-four or more years ago that's still applicable to Radio Amateurs today. That is the point *ham radio* magazine was making in our December "Reflections" editorial: reflections of past achievements that should be remembered, used, and built upon.

A few years ago I purchased a frequency meter LM14-Type 74028, built by Loral Electronics. Unlike the normal BC221 frequency meter, this unit uses one each of valve types 77, 6A7, and 76. For some time now I have been wishing to convert the unit to solid state using fets and have commenced doing so, attempting to adapt modifications to the BC221 type which uses two 6SJ7 and one K8-type valve. However, I have run into trouble in getting the variable oscillator to oscillate on the high band and am still attempting to get the crystal to oscillate.

No doubt over the years several ham radio operators have converted the LM series with this old type of valve and I write to you to see whether any such circuit has been published in your magazine. — Harry Hendriks, VK2ZHX.

One Amateur has reworked the same unit you have. Floyd K. Peck, K6SNO, in his article "More on Solid-State Conversion of BC-221/LM Frequency Meters," provides a solution replete with circuit and mechanical modifications. It appeared in the December, 1979, issue of *QST* on page 59. Floyd modernized his BC-221 using a 2N3819 for the solid-state VFO and MPF102 for the modulation mixer. In addition, he mounted the conversion parts in the bases of old 5-, 6-, and 7-pin tubes.

I am using an X-band mixer in a configuration that includes a 1N23WE diode followed by a 2N5179 i-f amplifier. The best noise figure I am able to achieve is 12 dB. Do you know of a lower noise diode in the same package as the venerable 1N23? Would a new Schottky barrier diode be a direct replacement? — Hilary McDonald, Jr., N5AX.

Several possibilities exist, but first it's necessary to mention that low-noise operation requires that the diode work into a well-matched circuit and that it be followed by a low-noise i-f amplifier. Lack of either might be contributing toward your 12 dB noise figure.

Specifically Microwave Associates specs their 1N23WE diode at 7.5 dB and their 1N23WG at 6.5 dB when operated into a well-matched circuit followed by 1.5 dB NF i-f amplifier. Finally, Hewlett Packard's Microwave Semiconductor Division specifies their Schottky barrier diode #5082-2713 at 6.0 dB under the same operating conditions and circuit. It is sold as a double-prong, type-49 package for $32.00 in quantities of 1 to 9.

Thanks go to Jack Lepoff at HP's Microwave Semiconductor Division and John Minck at the Stanford Park Division for the information.

ham radio

graphic azimuth and elevation calculator

A quick and accurate way to locate the geostationary satellites without a calculator

After I spent several weeks calculating the azimuth and elevation for the GOES series of weather satellites in many different locations, I decided there had to be a better way to do it than punching out the program over and over on my calculator. With a few moments of thought and an idea from a Meteostat (the European weather satellite) newsletter, I sat down to draw a simple graph which could be used to estimate the azimuth and elevation of ground stations for any of the geostationary satellites quickly. Since all the geostationary satellites are in the same orbits (except for their longitude), the only variable on the graph from satellite to satellite is the longitude difference between the ground station and satellite.

Many articles have been written describing the calculations needed to track the low orbit satellites (such as the OSCAR series).[1,2,3] The most popular of these is the Oscarlocator[4] graphical technique. This method is popular because the clear plastic overlay lets the operator quickly estimate the satellite location without lengthy computations. Even though minicomputers are rapidly invading the ham shack, few of us have been able to implement the real-time programs which show satellite azimuth and elevation as the satellite passes overhead.

simple geostationary calculations

All these complicated calculations are avoided with the geostationary satellite. These satellites orbit the earth 23,000 miles from the equator and make one orbit in exactly 24 hours. This keeps the satellite over a fixed spot on the earth. To the earth-bound observer the satellite appears fixed in the sky. Thus, once the antenna is fixed on the satellite no further antenna pointing is required. This is especially useful for unattended earth stations for TV and communications.

By Noel J. Petit, WBØVGI, 725 E. 51st Street, Minneapolis, Minnesota 55417

The major disadvantage of the geostationary satellite is that the tremendous distance between the earth and the satellite requires high-power transmitters and sensitive receiver systems to communicate reliably. Most satellite ground stations use parabolic dish antennas to receive the UHF and microwave signals from the satellites. Parabolic dish and other high-gain antennas have narrow beam widths, on the order of a few degrees. To estimate the width of the antenna beam use the following expression:

$$\text{width (in degrees from half power point to half power point)} = \frac{70^\circ \times \text{wavelength}}{\text{antenna diameter}}$$

Both the wavelength and antenna diameter must be in the same units. For example, both must be in centimeters. The beam width for a 2-meter diameter dish at 1691 MHz (17.74 cm wavelength) would be:

$$W = \frac{70^\circ \times 17.74\ cm}{200\ cm} = 6.2^\circ$$

This is a fairly large beam width; with such a large width, the calculation of the antenna azimuth and elevation need not be very accurate to initially find the satellite in the sky. Even a 3-meter diameter dish at 4.0 GHz (7.5 cm wavelength), where the present TV satellites transmit, has a 1.75° beam width. Knowing the azimuth and elevation within one degree is sufficient to find the satellite and then the signal-strength meter can be used to accurately align the antenna.

A graphical means of estimating the azimuth and elevation to a stationary satellite should therefore be accurate enough for most Amateur applications. **Fig. 1** is such a graph. The straight vertical and horizontal lines on the chart are lines of longitude difference and latitude, respectively. The curved lines radiating from the center of the chart are lines of azimuth to the satellite in 10° increments. The curved lines concentric to the center of the chart are lines of elevation to the satellite.

elevation, azimuth

Elevation is the angle from the horizon to the satellite. If the satellite is directly overhead, the elevation would be 90°. Elevation runs from 0° at the horizon to 90° directly overhead (the zenith in astronomical terms).

Azimuth is a bit more complex. Azimuth is the angle measured in degrees clockwise from true north. Looking north you are looking along the 0° line of azimuth. Looking east you are looking along the 90° line of azimuth and so on. From the 0° azimuth at north the measurement goes the full 360° around to north again; therefore, west is 270° azimuth, for example. All azimuth measurements to stationary satellites from the northern hemisphere will be between 90° and 270° since all the satellites lie to the south.

With the meaning of azimuth and elevation in mind, look at **fig. 1**. The first number you need is the ground station's latitude. Locate this latitude on the chart. For purpose of example, let us use Minneapolis at 45°N latitude. This would fall somewhere on the horizontal line half way between 40° and 50°N in the upper half of the chart.

Next, determine the longitude of the satellite and the ground station. I'll choose GOES-Central for my satellite example, at 107°W longitude. Minneapolis is located at 91°W longitude. Thus, the satellite is 16° west of the ground station. On **fig. 1** find the vertical position of the −16° longitude line. If the satellite is west of the ground station use the negative side of the chart and if the satellite is east of the ground station, use the positive side of the chart.

The correct location of Minneapolis is noted on the chart with a cross (+). The elevation of the antenna can be interpolated by noting that the + is between the 40° and 30° elevation rings — a little closer to the 40° ring than the 30° ring. Thus, the elevation of the antenna should be about 36°. The + is close to the 200° azimuth line — about one-fifth of the way to the 210° line, so the azimuth of the antenna would be about 202°.

The actual calculation for the GOES-Central satellite is *elevation* = *35.77°* and *azimuth* = *202.06°*, from a ground station in Minneapolis. The graph

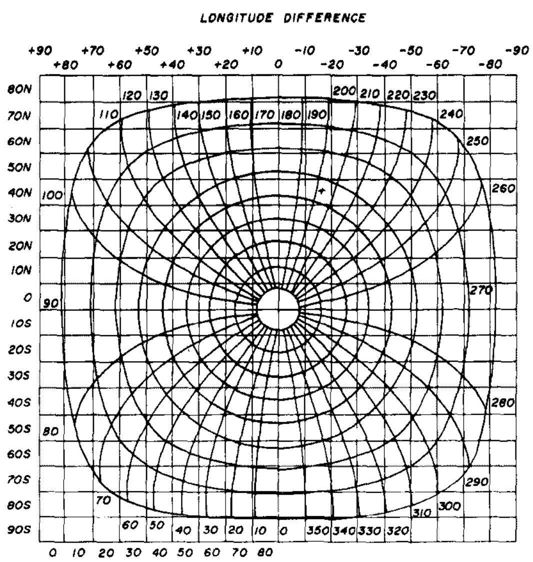


fig. 1. Azimuth and elevation calculation graph for geostationary satellites.

table 1. Geostationary weather satellite status as of January 1, 1982.

spacecraft designation	longitude	status
Meteosat	0°	operational — controlled by the European Space Agency
GOES-5	75°W	operational — GOES-East
GOES-2	107°W	operational — GOES-Central
GOES-1	127°W	standby
GOES-3	130°W	standby
SMS-1	131°W	standby
GOES-4	135°W	operational — GOES-West
GMS	140°E	operational — controlled by the Japan Space Agency

technique is accurate enough to use for most small antennas, but its big advantage is that there is little chance of making an error in calculation with this method.

A fairly complete list of the geostationary satellites can be found in a recent issue of *ham radio.*[5] However, the GOES-series of satellites is not accurately presented. Three GOES satellites are operational at any time. There may be any number of standby satellites in orbit in case the operational satellites experience a failure. Each satellite has two names (much like the naming system chosen for the OSCAR satellites). One name (like GOES-2 or SMS-2) designates the piece of hardware from the development stage to the time it was put in orbit. The other name designates the satellite when it is in use in space. **Table 1** lists the present status of the geostationary weather satellites in orbit as of January 1, 1982. This list includes the Japanese and European satellites which function much like the GOES satellites.

With this simple chart you can easily find the stationary satellites in the sky and tune in on the world of space communications.

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3. P.C. Bunnell, WA6VJR, "Tracking Satellites in Elliptical Orbits," *ham radio*, March, 1981, page 46.
4. C.L. Harris, WB2CHO and J.P. Kleinman, "The Oscarlocator," *QST*, November, 1976, page 51.
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ham radio

an interview with Dr. Kenneth Davies

Dr. Davies of NOAA discusses magneto-ionic effects and their relevance to radio propagation

Dr. Kenneth Davies, one of the leading experts in the field of magneto-ionic studies, has provided guidance in that field since 1949. At present, in his capacity as chief of the Research Division of the Space Environment Laboratory of NOAA, he supervises a technical staff involved in solar-terrestrial (interplanetary, magnetospheric, ionospheric, and atmospheric) studies.

During the past thirty-three years Dr. Davies has worked on projects investigating high-latitude ionospheric absorption and the use of oblique-incidence, pulsed ionosondes. At the famous CRPL (Central Radio Propagation Laboratory), he pioneered the high-frequency Doppler technique used for studying transient ionospheric phenomena. In addition, he was the principal investigator of the NASA/NOAA ATS-6 radio beacon project, which involved placing a frequency-, amplitude-, and polarization-stable transmitter on a geostationary satellite. The ATS-6 space platform provided continuous measurements of the electron content of the ionosphere.

Dr. Davies, author or co-author of over one hundred publications, is perhaps best known for his book *Ionospheric Radio Propagation*, which has become a standard reference in the field. While his professional credentials are very impressive, they are surpassed by his congeniality, his concern, and his ability to explain complex phenomena in terms a layman can understand. He is truly a scientist of highest magnitude.

ham radio magazine is pleased to introduce the first in a series of technical interviews with experts in various fields of communications. In this issue we are presenting excerpts from a two-hour interview with Dr. Kenneth Davies of the National Oceanic and Atmospheric Administration in Boulder, Colorado. Dr. Davies answers questions regarding radio phenomena that affect all Radio Amateurs.

This series will continue, in later issues, with discussions with experts in other related fields of direct interest to hams. The comments of our readers are welcomed.

ham radio: I understand that your chief interest lies in magneto-ionic studies. What is that?

Dr. Davies: Magneto-ionic studies deal with the propagation of electro-magnetic waves, more specifically, radio waves in a plasma (ionized medium) under the influence of an external magnetic field, the magnetic field of the earth. When a radio wave enters the ionosphere it splits into two waves, an ordinary wave (O) and an extraordinary wave (X). The waves are reflected at different heights, travel with different speeds, and, most importantly, are absorbed differently. Generally, the extraordinary wave is absorbed more than the ordinary wave, especially near what is known as the gyrofrequency. (See glossary.)

hr: How does a radio wave leaving an antenna know whether it's going to end up as an O or X wave?

Davies: Basically, it doesn't know until it gets to the ionosphere. When a radio wave enters the ionosphere it moves the electrons there. The electrons bob up and down in reaction to the electric vector of the wave — very much like a cork in water when a wave passes by. They start oscillating. In the presence of a magnetic field they *can't* oscillate back and forth linearly. The electron motion is twisted by the

By Rich Rosen, K2RR, technical editor, *ham radio* magazine

action of the earth's magnetic field. On average, half the electrons rotate one way and half the other. The energy is divided roughly 50/50, but not always. There are certain circumstances in which the wave can become all X or O or you can intentionally polarize the wave to achieve this.

Suppose you were at the geomagnetic equator. If you pointed your antenna in a north-south direction all the energy would go into exciting the O wave because the electric vector is parallel to the magnetic field vector. If you try to move an electron along a magnetic field at that location it behaves as if there were no field. (*Ed Note:* $\bar{F} = q\bar{v} \times \bar{B}$ where $\bar{F}$ = *force on electron, zero in this case,* q = *charge,* $\bar{v}$ = *velocity vector* and $\bar{B}$ = *magnetic field vector — the cross product of two in-line vectors = O*).

Now, at or near the gyrofrequency (approximately 1.5 MHz), the extraordinary wave is very heavily damped. If you use a vertical radiator at installations near the equator with East-West and West-East propagation you will excite the X wave. (*Ed Note:* This is of particular importance to South American, African, and other 0° latitude hams operating on 160 meters.)

hr: It's important in Amateur Radio for example because a number of hams who are very interested in DXing use both horizontally and vertically polarized antenna systems.

Davies: That is what magneto-ionic theory is all about. Understanding fading is important if one is to understand radio wave behavior. The fact that signals fluctuate can be related to O and X ionospheric theory. The O and X waves travel by slightly different paths which change as the sun rises and sets. When the two waves beat, deep fading occurs (as at sunrise).

Fading can also result from wave reflection from different ionospheric layers, and sometimes even different reflections from the same layer. This can be explained in terms of layer movement producing phase differences (including 180 degrees) between components of the same wave.

satellite communications is affected by propagation phenomena

Davies: Trans-ionospheric propagation (propagation through the ionosphere as occurs in satellite to ground communications) provides another example of O and X wave phenomena. A linearly polarized wave emitted from a satellite is decomposed into O and X waves that travel at slightly different velocities and increasing phase separation. The combined wave of constructively interfered components is equivalent to a rotating electric vector (changing polarization). The ionosphere changes both daily and seasonally, and the signal polarization with it. During the morning, for example, the electric vector may rotate in one direction and during the afternoon in the other. Over a long period of time conservation must occur; the net rotation must be zero.

hr: It sounds like the length of the rotation period might be related to some very basic phenomena.

Davies: Yes. If conditions on the sun and in the earth's atmosphere were stable then the polarization change — say from 00 hours one night to 00 hours the next — would not exist. In practice, that doesn't happen. The reason for this is partly that the sun varies from day to day, and therefore conditions are different, and also that the atmosphere is different: The upper atmosphere is very much like the lower atmosphere; there are winds up there. There are irregularities. These winds vary and affect the electron density.

hr: Are we talking about a completely random process?

Davies: You know that's our tendency. People have a tendency to say that if you can't explain something it's random. The answer is that this looks random until you understand what is causing it. These conditions are not random in the sense that they fluctuate for a reason. The reason is that the sun varies. As ionized particles from the sun approach the earth they set up electric fields from dawn to dusk across what's called the magnetosphere, the outer region of the earth's atmosphere. This active field in turn affects the plasma in the ionosphere. So a varying sun causes, through several intermediate steps, ionospheric variations. If we knew more about the detailed physics of the sun, this process might no longer be considered "random."

hr: I differentiate between a process that is completely random and one that appears to be random but is really not — or is everything causal? That is, we just don't know the cause?

Davies: This is my point. I would say that everything that happens in the ionosphere, as far as we know, happens for a definable reason.

hr: I'm thinking of the fact that you can build what's called a random-number generator. That is truly random. Or is that causal as well?

Davies: One source of random signals produced electrically is just a hot wire: the random motion of electrons in a resistor. But again it's random in the sense that there are a very large number of them, all of them moving in different directions. We cannot specify what any one electron is doing. You can specify the aggregate changes but you can't specify one.

solar activity indicators

hr: What are the best indicators of solar activity? I

understand that ultraviolet, as a measurement technique, is preferred over the 10.7-cm flux readings (see glossary).

Davies: An indicator of solar activity is the sunspot number. Sunspots, darker and cooler areas on the sun, were discovered by Galileo about 1610, and measurements of them have been made more or less continuously since the early 18th century, about 1715. These numbers show distinct cycles, the most prominent of which is the eleven-year cycle. There are longer cycles and there are shorter variations. For example a shorter variation of twenty-seven days is related to the rotation of the sun, as seen from the earth.

What we call the sunspot number is determined by a simple formula that gives a lot of weight to groups of sunspots:

$$R = K(10g + s)$$

where R = sunspot number
K = correction factor
g = number of groups of sunspots
S = number of individual sunspots

The question is, How good an index is this? Obviously, two different observers might see different groups on the sun. One observer might say that two groups exist — another observer that it's only one big group. Consequently any individual value of sunspot number is subjective. But if you have a sufficient number of observers over a long enough period of time, then you obtain an index which is at least roughly independent of individual preference; that is, the various biases cancel out.

hr: Why 10X (weighting) for a group (of sunspots)?

Davies: I don't know the exact historical reason for this, but it has to depend on a belief by solar astronomers (not ionospheric) that groups are more important than individual spots. It has been found that the critical frequencies of the ionospheric layers vary roughly linearly with sunspot number. This is important since sunspot numbers enabled people, in the early days, to project into the future. They could see a long series of cycles and predict what the next one was going to be.

Systems have been developed for predicting future sunspot cycles. One of these developed at CRPL is the McNish-Lincoln method. Data from an ascending cycle enables you to predict, fairly reasonably, sunspots for the next year or so.

hr: Could they have predicted cycle 21's demise based on its rise?

Davies: I'm not sure how long they took to predict the peak, but the prediction was very good for this cycle once it started. Now there are departures. You know this was a steep cycle. And there are differences, perhaps ten or twenty units. However, in general there is close correspondence between the predicted and measured curves.

hr: Cycle 21 (present cycle) had a twelve-month maximum moving average of a little over 160. A very nice cycle. And this one, what a beauty (looking at cycle 19) with 201 smoothed sunspot numbers.

Davies: That's the grand-daddy of them all.

hr: It was an interesting time to be an active Radio Amateur.

Davies: That's right, 1958. (Cycle 19 was the largest modern recorded sunspot cycle, peaking late '57, early '58.)

hr: What about ultraviolet as a measurement tool?

Davies: Let's go back to sunspots for a moment. Why should sunspots be related to anything in the ionosphere? They are just black patches on the sun. To answer the question, let's look at how the ionosphere is formed. The F-region of the ionosphere is caused mostly by solar radiation in the region of 300 to 1200 Å (Angstrom unit = 10^{-10} meters). The EUV (Extreme Ultra-Violet) of the sun's radiation ionizes the oxygen in the upper atmosphere. But you can't measure that from the ground, since it's all absorbed in the ionosphere. If you look at these radiations from satellites above the ionosphere, you find that there's a high correlation between them.

hr: Which satellites? You mean satellites in geostationary orbit?

Davies: Orbiting satellites appreciably above the maximum of the F-layer.

X-rays, on the other hand, having wavelengths of roughly 1-10 Å, come through the F-region. They have much more penetrating power and ionize the D and E-regions. An important point (hams take note!) is that over a sunspot cycle the EUV can increase by a factor of perhaps ten from low to high sunspot number, whereas the X-ray flux may increase by a hundred or a thousandfold during that same period.

hr: No wonder my 3.8-MHz signals get so absorbed during the sunspot cycle peak.

Davies: There's a much bigger solar factor variation of X-rays than there is of EUV. As you go further into the longwave part of the spectrum this becomes even more noticeable. For example, in the visible part of the spectrum there's hardly any change. The solar constant, the amount of energy falling on one square meter, doesn't change very much, less than approximately ½ to 1 percent.

Solar flares are another important category of emissions from the sun that affect terrestrial conditions. They are shorter-term fluctuations lasting from several minutes to days. A sudden ionospheric disturbance, a direct effect of X-rays from flares, can cause a radio blackout. The D-region experiences a tenfold increase in ionization. The D-region is important because, even though there are fewer electrons there than in the F-region, there are many *collisions* with neutral molecules. Remember that when a radio wave enters the ionosphere, the electric vector sets electrons in motion and, normally, if there were no collisions that energy would be restored, since a moving electron radiates. However, if the electron hits a molecule the energy that's re-radiated is basically converted into heat (from the motion) and radio noise; that is, random, electro-magnetic noise is generated. Once a collision takes place, the electron loses energy, and in the D-region these collisions occur often. This is the region where the energy is absorbed from the radio wave.

hr: What about the 10.7-cm flux?

Davies: The 10.7-cm flux technique is a good measurement. It has several advantages. First, unlike the sunspot number, it is a measurable quantity. Secondly, it penetrates the earth's atmosphere and can be measured on the ground. You don't have to see the sun. On a cloudy day it can still be measured.

hr: How accurately can you measure this?

Davies: It can be measured to roughly 0.1 percent.

hr: One thing still confuses me. It's possible in approaching cycle 21 minimum to have an SSN (twelve-month moving average) of 20 or 30 or less and yet on any particular day to have a high solar flux indication. Which should a communicator follow more closely, the SSN or flux measurements?

Davies: I would think that the 10.7-cm flux is probably the best index because of the reasons I've already given.

hr: Is it possible to have a fairly heavily ionized ionosphere even during a time of solar minimum?

Davies: Yes it is. Even at a sunspot minimum, severe magnetic and ionospheric storms can occur. In speaking of storms, by the way, these are disturbances produced by particles impinging on the earth that affect electrons stored around the earth. When a disturbance occurs in the solar wind, these particles accumulate near the earth, accelerate, and plunge into the earth's atmosphere. They are stored in the magnetosphere, the Van Allen radiation belt. These storms are very important from a radio communications point of view because they are the most destructive of all the ionospheric effects. They can last for two or three days and cause the critical frequencies (maximum frequencies) to drop by a factor of two or three, which means the spectrum available to communicators is reduced. This is particularly noticeable in high latitudes, near the auroral zone, because that's where the particles are.

hr: Does it enhance PCA (polar cap absorption)?

Davies: Yes, generally, but not always. Polar cap absorption events are due to energetic protons over the polar caps — but *inside* the auroral ring. Whereas storms *start* in the auroral zones. Storms are much more important than PCAs.

PCAs are relatively rare. You get a few per solar cycle and they're confined to high latitudes where the density of population is low. Some people mistakenly label an ionospheric storm as a PCA when they really mean ionospheric storms. Ionospheric storms are due to the arrival of particles of lower energy; not millions of electron volts, but a few thousand EV.

In a severe ionospheric storm energy is dumped into the auroral zone but the effects are carried by atmospheric circulation to low and middle latitudes where they can play havoc with radio communications by depressing the upper frequencies.

hr: I saw a report of a storm's effects lasting up to fourteen days.

Davies: That's right, especially in the lower ionosphere. It's called the storm after-effect and disturbs very-low-frequency navigation systems like Omega. Although the storm has died away as far as magnetic disturbances are concerned, the storms after-effects persist in the ionosphere.

cycle 21 characteristics

hr: We have talked briefly about cycle 21. Is it more or less typical, or has it surprised people?

Davies: It's taken many people by surprise. And the reason is this: Although it's a relatively big cycle with respect to ionization, so far as disturbances and storms are concerned it's relatively modest. It's not what we would have expected. Until this year, when we have had some disturbances, the number of storm effects have been small given the size of the cycle. The reason we're surprised, of course, is really that we don't know much about the solar cycle; that is, our sample of previous solar cycles is relatively limited. That is why we need to continue observations of the sun and related geophysical effects.

hr: Where are we in the cycle right now?

Davies: We're about halfway down the declining part of the cycle.

hr: What would our smoothed sunspot twelve-

month moving average be?

Davies: The latest sunspot number for June, 1982, was 110.

hr: That's still very high.

Davies: But then the month before it was down to 81. Whereas the month before that it was up to 122. Yes it is high.

hr: For January, 1983, for example, which is when this information will be published, what do you estimate?

Davies: In January we can probably expect numbers of the order of 90.

hr: Do most cycles go down to 5 or 10 (SSN) regardless of their maxima? Cycle 19, for example?

Davies: Yes roughly. They go down to about 10, some go to 5. Big cycles can be followed by small minima. Cycle 18, which began in 1944, was a big cycle with a very low minimum.

hr: How do people predict future cycles based on past cycles?

Davies: There are a number of ways. One way is to take a time series analysis of previous cycles, reducing it into simpler terms or waves with different periods. Then predict the future by getting equations for each of these waves. This is a purely mathematical approach.

hr: An orthogonal series like a Taylor or Fourier?

Davies: Fourier usually, or a new method, developed in the NOAA laboratory by Dr. Paul, called anharmonic frequency analysis. It provided a very good estimate for cycle 21. But, depending on what prediction method you use, you can come up with very different cycles.

hr: If, to predict cycle 21, you were to take all the previous cycles up to 21 and then do your Fourier or anharmonic frequency analysis, what would be the results?

Davies: It turned out that the maximum sunspot numbers predicted by that technique did, in fact, prove to be around 160 (cycle 21).

We're still a long way from an understanding of the sun, which is what we need to be able to predict. You must remember that all these methods tend to be recipes, not true understanding.

hr: Will we ever be able to accurately predict cycles? Let's look into the future. Is there any reason to believe that we will understand all the processes both quantitatively and qualitatively well enough to be able to say, "Cycle 475 is going to have a peak of . . ."?

Davies: Yes, I think that is possible in principle. After all, think of astronomy and how we can predict satellite and planetary orbits with extremely good accuracy. There's reason to hope that, given enough effort, the secret will be unlocked. At the moment we don't know if we're doing the right thing. We have to wait for a Newton . . .

extraplanetary effects on earth

hr: We talked about sudden ionospheric disturbances. Are there other causatives, such as planetary conjunctions?

Davies: Yes, I think it was Mr. Nelson of RCA who hit upon a method by which he predicted disturbances on the basis of planetary alignment. He said that when the planets were lined up on one side of the sun storms would occur. The interesting thing is this: it seems that he predicted a few of these storms within a number of days. But statistically, these methods do not stand up to scrutiny. The question I have asked people who criticize this is the following: If you could predict the big ones, isn't that important — much more important than missing out on the small ones?

less well known ionospheric layers

hr: Are there layers other than D, E, and F?

Davies: There have been some claims for a C-layer which is formed about 60 km up, caused by cosmic rays and is therefore always present. There is another phenomenon seen on ionograms during a storm when the critical frequency of a regular F_2 is greatly diminshed: a little kink appears very high up and is called a G-layer. It's probably not a distinct layer but is a phenomenon that occurs, let's say, at special times only.

hr: Would it correspond to the normal height of the F-layer? Or is it even higher?

Davies: It's difficult to say how high it is because on an ionogram some of the apparent height is caused by retardation underneath. It's not a true height of reflection. The height is actually lower than the height you see on the ionogram.

hr: We're talking virtual heights on an ionogram?

Davies: Yes, virtual heights. During solar eclipses, in certain parts of the world, the F_1-layer tends to break up into F_1 and $F_{1\frac{1}{2}}$. Then the E-layer at times has little kinks in it called the E_1-layer and the E_2-layer. But you must remember that these are not necessarily physical phenomena. They are the appearances of traces on the ionogram.

hr: Are new techniques and instruments now available that will help us differentiate between an apparent and an actual height?

Davies: The new digital ionosondes (conventional sounders are analog) will help us interpret the results. What they will tell us is where the signals actually come from: whether they arrive from vertically above the sounder or off to one side. What the new sounders will therefore do is help us interpret the structure of the ionosphere. That will help.

We've found that some of these "extra" layers are not extra layers. They are layers off to one side at the same height — perhaps parts of another layer. (As an example, if you're 45 degrees off vertical and get a reflection from a hundred kilometers away, you might see another trace at 140 km. Without any other information it's possible to interpret that as a layer at 140 km.)

A greal deal of effort in the 1940s and 50s was put into interpreting echo traces on ionograms. Without full knowledge, it's a very frustrating exercise and in many cases fruitless.

A technique that gives the true height of the ionospheric layer directly is the incoherent scatter technique, in which radio power is scattered by individual electrons. This requires very large and expensive systems, and there are only about six of them in operation.

modes of propagation

hr: Most Radio Amateurs are familiar with the terms one-hop, two-hop via the E or F-layer, but what about M-type: $F_2 - E_S - F_2$ and some of the lesser known modes?

Davies: Well, for example, in trans-equatorial propagation, there's evidence that you can get, because of distortion of the ionosphere, an equatorial anomaly. You can get two reflections off the ionosphere without an intermediate ground reflection. It happens in the evening hours around the equator. There are relatively few of these mechanisms of propagation that are really well documented and well founded.

There have been suggestions that transequatorial propagation is really due to ducting — ducts which go way out, well above the ionosphere, and trap energy.

hr: Well above the ionosphere?

Davies: Yes, or into the upper parts of the ionosphere.

hr: When you say well above, what are you referring to? 500 km?

Davies: Maybe 500, 1000, or 2000 km. But it's not as great as Whistler waves, which are VLF waves, audio-frequency-type waves, which can go out to very large distances, like 30,000 km.

long delayed echoes

hr: It is said that there's a phenomenon called LDE (long-delayed-echo), approaching a second. Do you know of any reason for an LDE that long?

Davies: It's not a field I've closely followed. Plasma physicists say that there is some instability — a storage in the plasma — that takes a long time to release. The people who have studied it insist that these echoes are real. They have coded the signals and claim that they have received real long-delayed returns.

Maunder minimum

hr: Is there evidence today to indicate that the Maunder minimum actually did occur, or is it felt that it was merely a result of lack of recording capability?

Davies: I think that there is evidence that there was a Maunder minimum. It's not likely to have been simply a lack of data.

hr: Is it probable that future generations will see a Maunder minimum?

Davies: I would say yes. It's certainly quite possible. It could happen many times in fact. It could happen the next cycle. Who knows what the next cycle is going to be?

sidescatter

hr: Can you explain the reason for signals sometimes deviating from great circle paths? One example that I recall is the U.S./Europe path via North Africa. Is this a well-known, often-occurring condition?

Davies: Yes, I think it is. It's been known for many years that signals from Europe often appear to come from directions roughly in the mid Atlantic or North Africa, the Sahara region. One suggestion has been made that this occurs when the direct path say from Europe to North America passes through a disturbed auroral zone. If that zone is disturbed, the direct path through the auroral zone is cut out. But you still receive a signal across the Atlantic. It is reflected off the coast of Africa; that is, it hops down to Africa, scatters there, and then across the Atlantic. That bent path is too far south for the storm to affect it. That has been documented and is a possible explanation.

hr: Are there other explanations for sidescatter?

Davies: I think that's the primary one. The other possibility for a signal coming off the great circle path

is ionospheric irregularities, particularly field-aligned ones. Irregularities in the ionosphere tend to be field-aligned because electrons can move easily along the field but not across it, creating sheets or columns of ionization. If the geometry is right you get perpendicular reflection.

twilight-zone propagation

hr: What is twilight zone (gray-line) propagation? It's a term Radio Amateurs use to explain a non-great-circle path that provides an unexpected opening.

Davies: I know that people claim RTW echoes are signals received long distances along the twilight zone, or earth's day/night dividing line. It seems to be some form of ducting.

forecasting and predicting

hr: Can you explain the difference between the terms forecasting and predicting?

Davies: Yes. When people talk about predicting, they mean long-term: predicting what the critical frequencies will be next year or the year after or during the next sunspot cycle. This is done for long-range planning purposes: what frequencies you should request, what antenna should you design, angles of takeoff, arrival, and so on. Propagation conditions for the long term are usually referred to as predictions.

Forecasting on the other hand is usually thought as short term, in terms of solar flares.

forecasting solar flares

Davies: To forecast solar flares, we must keep a watch on the sun all around the world and report what happens when a solar flare has occurred by optically viewing the shape and features of the sun. These are solar features such as plages, magnetic configurations, etc. We measure the way the solar magnetic fields become distorted. It's an energy build-up at these active regions of the sun. When this occurs we also receive other information from the sun, solar radio emissions for example. One way of forecasting a PCA is by observing the shape of the spectrum, the so-called U-spectrum wherein the flux has a U (or dip) in the microwave region. There's a minimum in the spectral distribution.
HR: Just prior to the disturbance?

Davies: Yes. We look at the sun in visible light and X-rays. X-rays are very important. X-rays are classified C(s), M(s), and X(s), in which an X flare is a large flare which causes short-wave fadeout (the system was developed in the Space Environmental Services Center). This is one of the best ways we now have of measuring the size of a flare and its ionospheric effects. It's possible to have a big flare optically that may have almost no effect on the earth.

hr: Is this because it misses the earth?

Davies: No. It may be because the spectral content is different. Remember you can have two large flares: one has strong X-ray burst components that ionize the D-region and cause blackout; the other flare, which looks the same in visible light, may not have the X-ray components.

Also, the other point you've raised is valid. The earth must be in the proper position to be hit by the particle burst. We don't know at the moment what the relationship is between the visible appearance of a flare and its particle output. That is something that we must determine.

future of forecasting

hr: Let's look ahead twenty years. Do you see a considerable improvement, based on the work that is occurring right now in the lab, in terms of forecasting?

Davies: Well, unfortunately no. The outlook is very bleak I must admit. Under the President's budget all research, for example, in this lab, Space Environmental Research, will be abolished.

tools of the trade

hr: Can you briefly describe the function of the following instruments: auroral radar, Faraday polarimeter, digital ionosonde?

Davies: An auroral radar is a fixed-frequency radar which bounces signals off ionization in the auroral zone.

A Faraday polarimeter is a device that measures the rotation of the electric vector of a wave and, hence, the electron content of the ionosphere.

A digital ionosonde is a recording instrument that provides standard amplitude and height information (as in conventional ionosondes) and parameters that can be used to determine polarization, phase, and direction of arrival of the signals.

hr: Can't these ionosonde tasks be accomplished using conventional instrumentation?

Davies: Yes, separately. The new digital sounders enable all of these to be done together. That's the big advantage of the new system.

The major thing that we are able to do with new information is determine where the signals come from — and therefore what the horizontal gradients in the ionosphere are. The ionosphere is neither a flat nor a concentric medium.

For example, there's a mid-latitude trough (depression in electron density) in the F-region just below the auroral zone. It's very important, since it distorts the ray path of over-the-horizon radars.

interest in hf

hr: This naturally brings us to my next question. Why is there renewed interest in hf?

Davies: It's related to the obvious advantages of hf. The installations cost less and the system is useful for long distances. There was a feeling in the 60s, when geostationary satellites were beginning to be used, that our radio communications problems had been solved. We thought we could get rid of ionospheric effects by going to satellite communications — the ionosphere would have no effect. But the ionosphere does have an effect, even at these frequencies, in the form of scintillation; even up to 6 GHz. Also, of course, from a vulnerability point of view, satellites are perhaps more vulnerable than the ionosphere to an enemy action. You can wipe out a satellite, but it's more difficult to wipe out the ionosphere.

hr: You mentioned that the ionosphere is extremely forgiving.

Davies: Yes, I think this is true of the entire earth. It recovers from abuse. We pollute the ionosphere intentionally and unintentionally. Huge electron depletions have been noted during the launches of Saturn rockets for example. Thousands of kilometers are affected.

hr: Temporarily?

Davies: Yes, perhaps for an hour or two at night. The sun replenishes it. Even though huge quantities of O_2, H_2, and water are put into the atmosphere, these are natural constituents. They don't do long-term damage.

future of hf

hr: What's the future for the high frequencies?

Davies: I think they will always be with us. I can remember when I started my professional career, people said to me the high frequencies are dead. That was thirty years ago.

glossary

10.7-centimeter flux — Solar radio noise flux at 2800 MHz that is indicative of the degree of E-layer ionization.

Maunder minimum — A name applied to a period of approximately 70 years in the late seventeenth and early eighteenth centuries, during which solar activity is believed to have been unusually low.

Gyrofrequency — Gyrofrequency, also called the gyromagnetic frequency, is a frequency derived by the formula below that describes the motion of an electron in a magnetic field. In the ionosphere, where the magnetic field intensity is roughly 0.5 gauss, the gyrofrequency is approximately 1.5 MHz.

$$f_H = \frac{1}{2\pi} e/m\, B$$

where f_H = gyrofrequency
e = the charge on an electron
m = the mass of an electron
B = magnetic field intensity

ham radio

remote-site receivers
and repeater operation

A system for selecting the strongest signal from remote-site receivers

Repeaters are usually built for mobile and hand-held use. Usually the repeater transmitter has a 20 dB power advantage over a hand-held. At the same time, the received signal at the repeater is reduced by this same figure. This means we can hear the repeater transmitter much better than the repeater receiver hears the hand-held.

One of the easiest ways to equalize coverage is to install remote receivers (commonly referred to as satellites). This concept has been used successfully for many years on commercial fm systems, providing full quieting signals over an entire metropolitan area.

satellite operations

A remote receiver system places receivers in locations distant from the main repeater transmitter site. The remote receivers relay the weaker signals back to the transmitter via UHF links. The ideal system, shown in fig. **1**, uses three remote receivers feeding one transmitter. Since the receivers are located away from the transmitter, you can run *higher* transmit output power without desense.

So far this seems fairly simple; however, the big problem is designing an effective voting system where the transmitter site will vote (select) the strongest, fullest quieting signal. In commercial systems voting becomes very elaborate, almost impossible for Amateur repeater groups to maintain and afford. Motorola uses a system of encoding tones at various levels in which an elaborate gate system selects the receiver hearing the best signal.

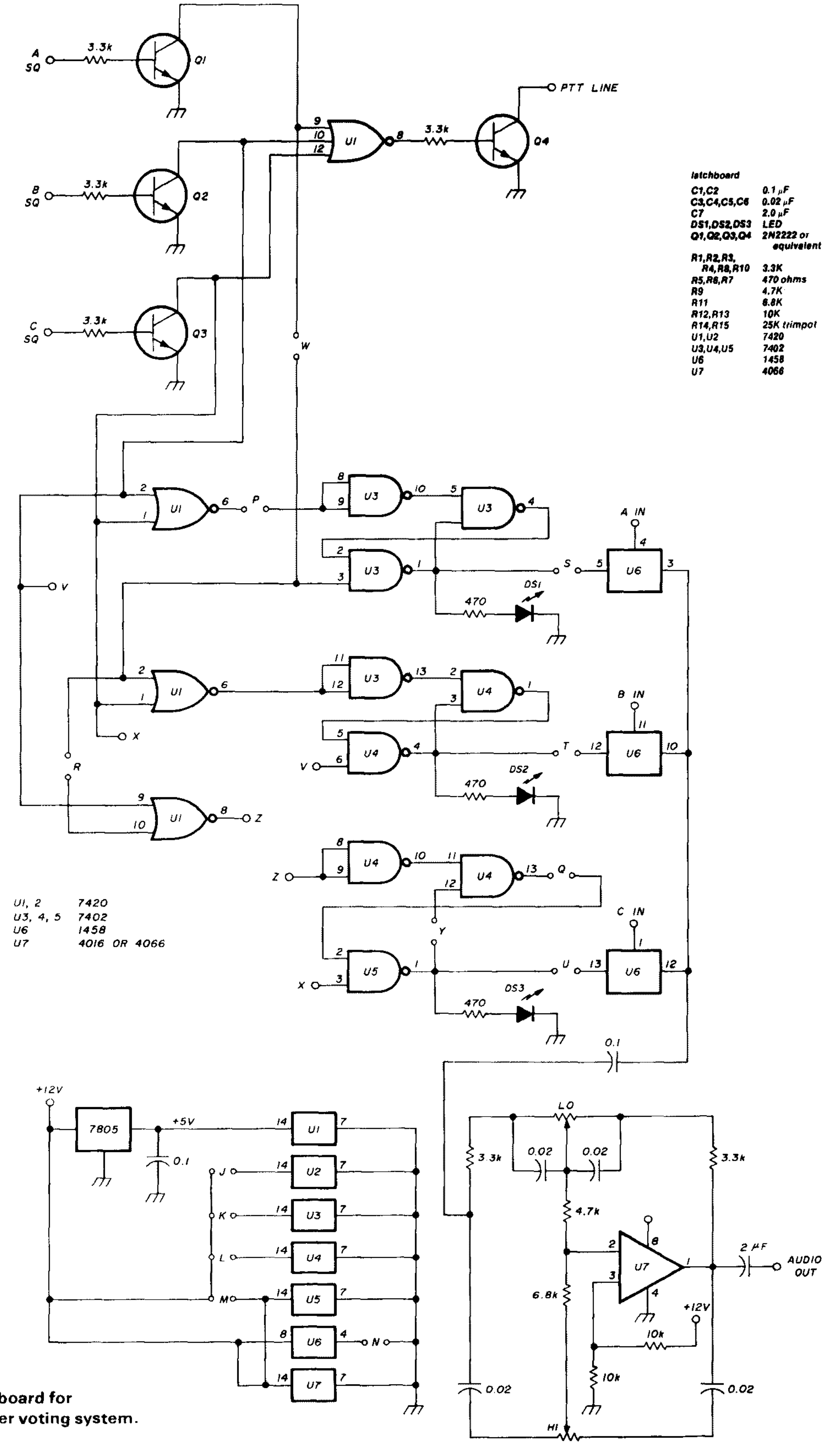
Many Amateur repeater groups could use multiple receiver site systems, however, to date there has not been much in the way of a simple and inexpensive voting system described in Amateur literature. This article describes a very simple voting system that will allow three remote receivers to feed one transmitter with the best quieting signal. It has served our Marissa, Illinois, repeater system very well.

The transmitter site contains three link-receiver boards which operate on the 220 MHz band. These boards were removed from Midland Model 13-509 220 MHz transceivers while the corresponding transmitters were used at the three receiver sites. At WD9GOE each of the 220 MHz receiver boards were mounted on a separate chassis with proper connectors to plug in to a main frame. This allows for easy installation and quick service should malfunctions occur. All three receivers are exactly alike except for the operating frequency. Each 220 MHz transmitter operates on separate frequencies. WD9GOE/R uses 30 kHz separation between the three link frequencies.

voting secret is time

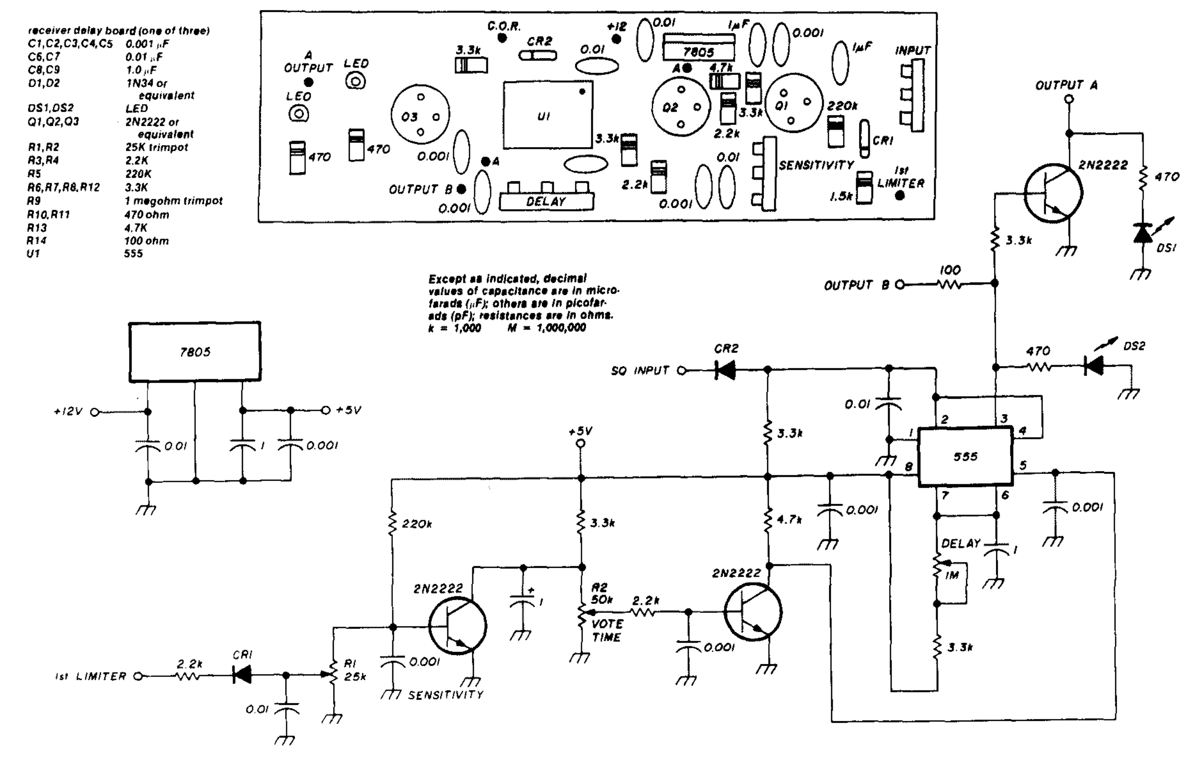
The secret to this simple voting system relies on the time it takes the 2-meter receiver to turn on the link transmitter. A full quieting signal turns the link on immediately. A signal that is − 10 dB quieting or

By Bob Heil, K9EID, #2 Heil Industrial Blvd., Marissa, Illinois 62257



latchboard	
C1,C2	0.1 μF
C3,C4,C5,C6	0.02 μF
C7	2.0 μF
DS1,DS2,DS3	LED
Q1,Q2,Q3,Q4	2N2222 or equivalent
R1,R2,R3, R4,R8,R10	3.3K
R5,R6,R7	470 ohms
R9	4.7K
R11	6.8K
R12,R13	10K
R14,R15	25K trimpot
U1,U2	7420
U3,U4,U5	7402
U6	1458
U7	4066

fig. 1. Latchboard for three-receiver voting system.



receiver delay board (one of three)	
C1,C2,C3,C4,C5	0.001 μF
C6,C7	0.01 μF
C8,C9	1.0 μF
D1,D2	1N34 or equivalent
DS1,DS2	LED
Q1,Q2,Q3	2N2222 or equivalent
R1,R2	25K trimpot
R3,R4	2.2K
R5	220K
R6,R7,R8,R12	3.3K
R9	1 megohm trimpot
R10,R11	470 ohm
R13	4.7K
R14	100 ohm
U1	555

fig. 2. Satellite receiver delay board.

less delays the link transmitter from turning on for up to 0.5 second. Therefore, the first signal the transmitter receives will be the best quieting. When using multiple receiver sites, it is best to adjust the squelch controls of each remote receiver a little tighter than normal.

latching logic

At the transmitter site, the three link receivers' squelch turns on a dc switch that feeds a series of gates. The gate series accepts the first signal and locks out the other two. This type of priority latching system allows a mobile system to key up more than one of the receiver sites and, as the signal jumps from one site to another, will automatically switch sites. The gates ultimately turn on the PTT transistor switch which keys the transmitter (**fig. 1**).

The audio from the three receivers is selected by a 4066 switch. It switches the audio very smoothly. An active audio equalizer which allows loss in frequency response or level during the link process to be made up provides the system with excellent audio quality. The audio mixer/equalizer is an active device that can provide plus or minus 12 dB of dynamic range between 300 and 2000 Hz.

receiver delay board

The delay board used at each two-meter receiver site uses one LM555 timer IC and three NPN transistors (**fig. 2**). The limiter voltage from the receiver is fed to the VCO input (pin 5), and the squelch line is fed to the pin 2 control input. If the limiter current is high, meaning the receiver is hearing a very strong signal, the 555 turns on immediately. As the limiter voltage drops, which means a weaker signal is being heard, the 555 will delay turning on for 0.5 second or more. This delay can be adjusted with the 1M pot connected with pin 7 of the 555.

Alignment of the delay board is simple. Read the voltage at the pin 5 test point. Set R1 fully clockwise and R2 fully counterclockwise. Set R3 for a one-second key-up delay. With no signal input, pin 5 will be about 1 volt below VCC. Apply a full quieting signal to the receiver and adjust R2 until pin 5 reads 0.025 volt. Still applying signal, adjust R1 to decrease sensitivity. Decrease R1 enough to raise pin 5 by 0.2 volts, giving a reading at pin 5 at about 0.45 volt. With no signal, pin 5 should read about 3.5 volts. This level is dependent upon the value of VCC. Adjust R3 for the desired delay.

The output keying transistor of the delay board will key the 220 link transmitter so no other COR or logic is necessary at the receiver site. The delay board is small enough to mount inside a Motorola or GE receiver strip.

The accuracy of the 555 timing sequence is critical. A normal LM555 is useful from 0 degrees to 70 degrees C. Signetics makes a military spec LM555 which guarantees operation from −70 to +125 degrees C. Use of this device is highly recommended.

link benefit

This system gives one side benefit. Users of 220 equipment will be able to work two meters with their 220 gear by operating on one of the link frequencies, providing they can hear the link from the main receiver and the transmitter site receiver can hear them. Caution should be observed because there may be times a hand-held or mobile station may key a remote receiver no one is listening to. It is best to be able to monitor all three links to be safe.

link antenna system

The Marissa system operates their three receivers on a battery back-up system. Only 1 or 2 watts of link power is used and this feeds a seven-element Yagi. After much experimentation, we found that by using horizontal polarization for the link antenna system, we were able to improve the signal-to-noise ratio because most manmade noise is vertical in polarization.

battery warning beeper

Each receiver site has a warning beeper that activates when the site goes on battery power. One location high atop a grain silo loses commercial power at 5 p.m. each day and all day Sunday. A large tractor battery powers the site when the line voltage is off. To warn the control operators and tech crew that the site is on battery power, a beep sounds every 15 seconds when that receiver site is activated. The tone level is adjusted so it does not interfere with the received signal. Two 555 timers are used to do this (see **fig. 3**).

temperature problems

The first winter the system was used turned out to be a real challenge for the tech crew. The 220 rigs were built by Uniden in 1973 and 1974 and marketed under several names. At that time the marketing people thought the 220 band was going to be given to the CB service. The units were not designed to be used in extreme temperature changes. The biggest problems encountered when using them as link components is frequency shifts.

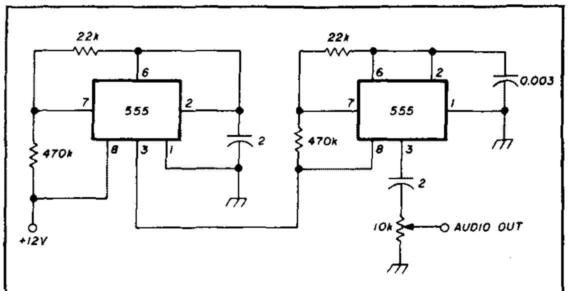


fig. 3. Emergency power indicator.

The problem is the TR-29 oscillator transistor circuit. It was found to be very unstable when operated below 30 degrees fahrenheit. A simple cure was to mount two small grain of wheat pilot lamps (removed from the front panel of the rig) as close as possible around TR-29. Hold them in place with silicone rubber compound. A 200-ohm resistor is used in series with the lamps to lengthen filament life. This was done around the crystal also. You could install a crystal oven but the pilot lamp trick works well and is much more economical.

WD9GOE/R

The system at Marissa has three receiver sites. One is thirty-five miles north of the repeater near Scott Air Force Base. The main site is five miles from the repeater transmitter, and the third, thirty-five miles south at Pickneyville. All receivers are Motorola Motran M series. The repeater transmitter runs 200 watts on 147.210 MHz using a General Electric CP-5 final with a 4CX250B. The system has been in operation for about three years and works very well. A similar system is in operation at Gillespie, Illinois, in WR9ACD/R, built by Jim Heyen. We have spent countless hours and shared many ideas on the design of this system.

It is normal to hear two-way hand-held communication across a 50 to 60 mile path. Mobiles enjoy solid 70 to 80 mile coverage. The system also has an elaborate link system with ten-meter fm, and six- and two-meter SSB. It also has voice identification and information tapes, time machine, autopatch, and several other features.

The remote voting receivers have helped to more than triple the coverage of our repeater system. They are inexpensive and require little, if any, maintenance, and will be a worthwhile addition to any system.

ham radio

The complete kit of drilled PC boards, all parts and instructions are available for $82.00 (plus $3 shipping) from Heil, Ltd., P.O. Box 68, Marissa, Illinois 62257. **Editor**

the simple simplex autopatch

Mix a transceiver, some parts and ingenuity to build this simple device

Many repeaters in the Amateur service employ a method of accessing the telephone system. Usefulness of this autopatch feature is reduced if the repeater is often used for normal communications. It is my experience that commercial autopatch systems are very costly and home-brew units can get quite complicated. A simpler and cheaper autopatch may be built around an available VHF/UHF transceiver by using a different method.

the simplex concept

The most expensive part of a repeater autopatch is the repeater itself. A significant cost reduction is realized by using simplex communication; transmitter and receiver do not have to operate simultaneously. This concept eliminates many problems since no antenna duplexer is needed, only one antenna is required, and a standard transceiver may be used.

The simplex concept has a problem: if the telephone side of the conversation is being transmitted by the patch, how does it determine when the primary user wants to talk? When should the system switch to receive?

One common design in commercial systems switches to receive whenever the person using the telephone stops talking. The person using the radio is now free to begin transmitting into the patch. This method is unsuitable in Amateur radio service since the ham must have complete control over all rf transmission, including the autopatch.

Another method of transmit/receive control is to switch the patch transceiver periodically into a brief receive window. The window checks the frequency for presence of a carrier, presumably signifying that the ham user wants to talk. If the carrier is present, the patch will remain in the receive mode until the carrier disappears. If no carrier is present during the receive window, the patch system immediately resumes transmitting until the next window test. No part of the telephone conversation is lost if these windows are brief.

A complete simplex autopatch system consists of four major components: the controller and telephone interface, the transceiver, a power supply, and an antenna. This article describes the controller in detail and shows how a complete system can be built around an ICOM IC-230 transceiver.

The controller is a dedicated unit and *not* microprocessor-based. I found that standard TTL provides a simpler control unit that is easy to trouble-shoot. Those of you without access to a PROM programmer will be relieved to know that this project does not require memory programming!

The patch circuitry is relatively simple but boasts such useful features as an automatic identifier, time-out circuitry, and full monitor capability. It can still be used as a normal transceiver even though the rig requires some squelch modification.

the controller

The control unit is the heart of the simple simplex autopatch. It performs the connect and disconnect functions between telephone and transceiver. It also generates the receive window for transmit/receive control and provides the CW identifier. Time-out circuitry and an audio monitor are included.

The controller is the small chassis on top of the IC-230 in **fig. 1** and operates from the same 12 Vdc supply as the transceiver. My system is powered by a battery to avoid problems when power is not available.

A block diagram of the controller is shown in **fig. 2**

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with the main schematic in **fig. 3**. Identifier, reset, and power supply sections are detailed in **fig. 4**. **Fig.** 5 shows the inside view of the controller chassis. Three tone decoders control telephone line connect and disconnect. The timeout section will switch the system to standby if the patch remains on for about two minutes with no incoming signal interruptions. The receive window generator section is described later.

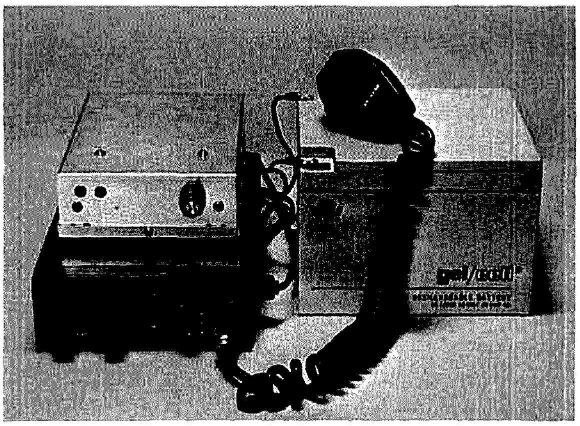

fig. 1. The Simplex Autopatch System. The autopatch controller is the small chassis on top of the ICOM IC-230 2-meter transceiver. The battery is optional and allows continued operation after power failure.

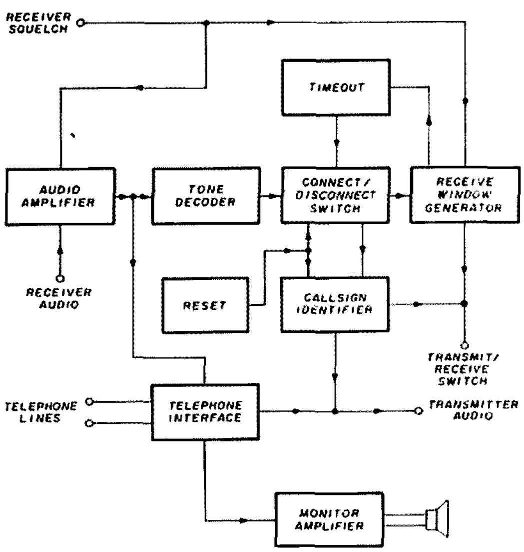

fig. 2. The autopatch controller circuit block diagram.

audio

Transceiver audio is taken from the receiver's discriminator so that audio into the controller is independent of the transceiver volume control setting. This low-level audio signal is amplified by U1 and sent to the telephone line through transmit/receive relay K2 and isolation transformers T1 and T2. Receiver audio also goes to monitor amplifier U2 and three tone decoders, U5 to U7.

Telephone line connection is from the red and green wires through line-connect relay K1. One line isolation transformer with 600 ohm impedance on each winding is normal. I used two Radio Shack units wired back-to-back, since they were inexpensive and easy to find.

The 567 tone decoder has a reputation for false triggering. I made an attempt to improve reliability in two ways: first, transistor Q1 turns off U1 whenever the receiver is squelched so that noise cannot activate the decoders. Second, the 567 switching speed is slowed by the 10 μF capacitor to pin 1 of each chip. The tone must be present for about one-third second to activate the decoders.

beginning the control sequence

U9A in **fig. 3** is a D-type flip-flop that holds the most recent tone command, either connect or disconnect. A connect command is applied directly to the preset input while the disconnect command resets U9A through its clock input, pin 3. The clear input at pin 1 is activated by the $\overline{\text{RESET}}$ (from **fig. 4**) or the time-out from U12. This configuration saves a gate since the $\overline{\text{RESET}}$ and disconnect signals, while performing the same function, are applied to different pins on the flip-flop. Gate conservation is important: every gate in any package is in use!

It should be noted that HI and LO markings in **fig. 3** are the TTL logic levels for the particular function. Audio amplitude levels are given later under control circuit alignment.

Flip-flop U9B is set each time a connect or disconnect command is decoded.* When either command disappears and the receiver squelch is on via inverter Q2, gate U4D starts the identification (ID) sequence by clearing U19B (**fig. 4**). U9B is cleared through D5 when the ID begins. This interconnection causes an automatic ID transmission at the beginning and end of each time period the patch is used. If each phone call is limited to ten minutes, identification requirements are fulfilled.

Telephone line connect relay K1 will pick up when both bases of Q6 and Q7 are pulled low. This occurs when the patch has been activated (U9A set) and the

*U8B and U8A will OR high inputs without inverting the output; when either output returns low, U4A output goes high to clock U9B on the positive-going edge. Editor

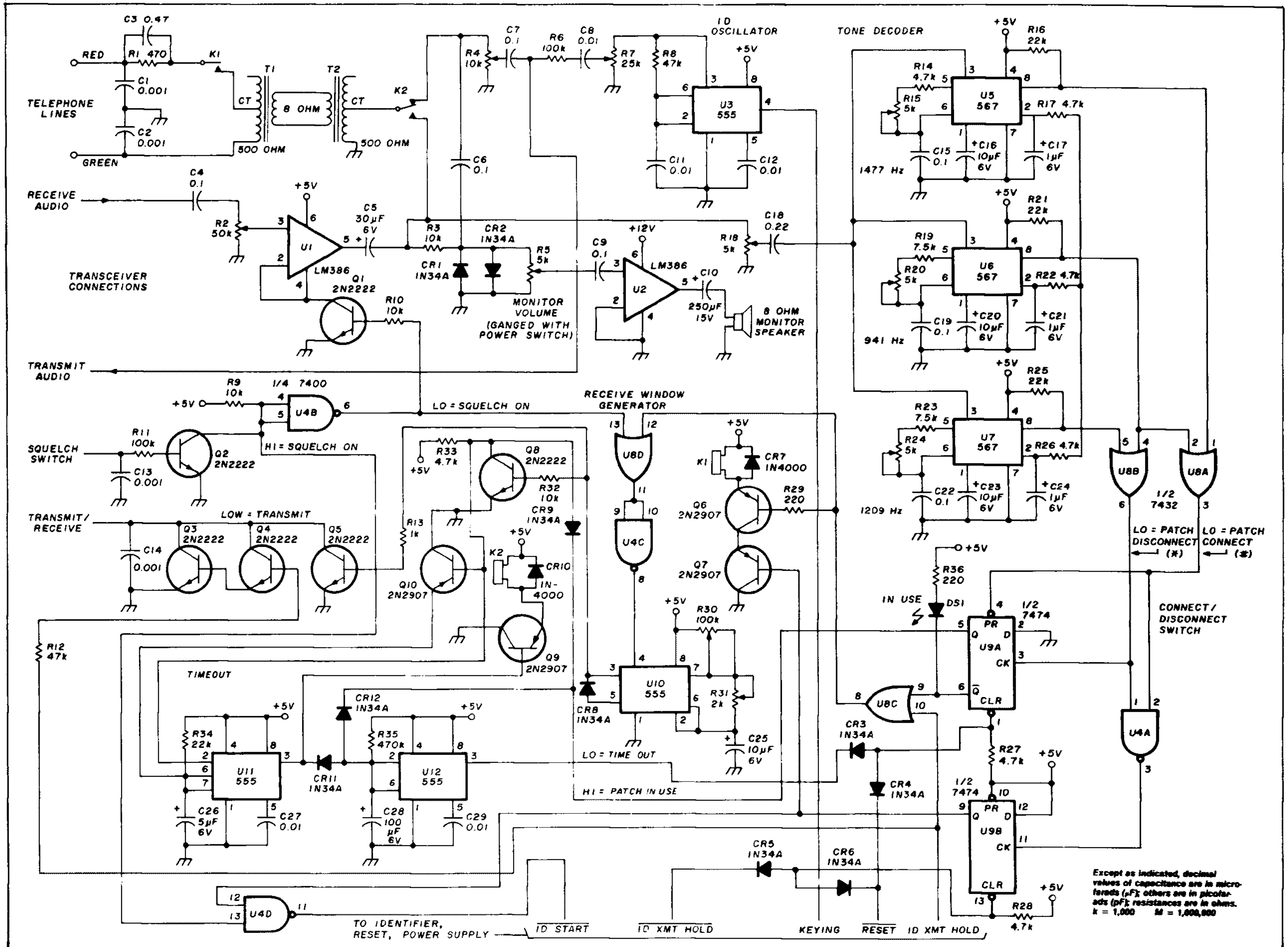

fig. 3. Autopatch controller main circuitry, including all interface connections to the transceiver and telephone line.

table 1. Parts List.

Resistors, 1/4 Watt, 10 percent unless noted	
R1	470, 1/2 Watt
R2	50K trimmer potentiometer
R3,R9,R10,R32	10K
R4	10K trimmer potentiometer
R5	5K panel-mount potentiometer
R7	25K trimmer potentiometer
R8,R12,R38	47K
R6,R11	100K
R13	1K
R14,R17,R22,R26, R27,R28,R33,R40,R41	4.7K
R15,R18,R20,R24	5K trimmer potentiometer
R16,R21,R25,R34	22K
R19,R23	7.5K
R29,R36	220
R30	100K trimmer potentiometer
R31	2K trimmer potentiometer
R37	680
R39	470K
Capacitors, μF, 15 V Disc unless noted	
C1,C2	.001, 250 V minimum, disc
C3	.47
C4,C6,C7,C9	0.1
C5	30, 6 V electrolytic
C8,C11,C12,C27, C29,C33,C35	.01
C13,C14	.001
C15,C19,C22	0.1 polystyrene or metal film, temperature stable, Radio Shack 272-1053
C16,C20,C23,C25	10, 6 V electrolytic
C17,C21,C24,C31, C32,C34	1, 6 V electrolytic
C18	0.22
C26	5, 6 V electrolytic
C28	100, 6 V electrolytic
C30	1, 15 V electrolytic
Transistors	
Q1,Q2,Q3,Q4, Q5,Q8	2N2222 or equivalent NPN
Q6,Q7,Q9,Q10	2N2907 or equivalent PNP
Integrated Circuits	
U1,U2	LM386N-1
U3,U10,U11,U12, U14,U17	LM555J
U4,U16	7400
U5,U6,U7	LM567N
U8	7432
U9	7474
U13	LM309K
U15	7473
U18	7493
U19	7476
U20,U21	74154
Diodes	
CR1 to CR6,CR8, CR9,CR11,CR12	1N34A or equivalent germanium
CR7,CR10	1N4000 or equivalent
DS1,DS2	16 mA LED, panel-mount
K1,K2	SPDT contacts, 5 Vdc coil, 56 ohms, Radio Shack 275-216
S1	SPST push button
S2	SPST 1 Ampere (ganged with R5
T1,T2	1000 ohm C.T.: 8 ohm, Radio Shack 273-1380

ID sequence has finished transmitting (U9B clear and U8C-10 low). K1 drops out on a disconnect command when U9A is cleared.

the receive window

Timer U10 is wired as an astable multivibrator, operating when the patch is activated and the receiver squelch is on (U10-4 high through U4C and U8D). When the timer output at pin 3 is high, it turns on the patch transmitter through Q5. The transmit/receive line is also held low for transmit during the ID sequence via Q4 and Q3.

If the squelch is tripped during the window provided by the low period of U10-3, reset pin 4 is brought low and turns off timer U10. The receiver will remain on. R30 adjusts the length of the receive window and R31 sets the time between windows. The diode on pins 3 and 5 of U10 ensures that the first pulse of the timer is the same length as all subsequent pulses.

Note that this circuit requires the transceiver to provide a low voltage level for receiver squelch and that the transmit control line should also be at a low voltage level. If your transceiver requires opposite levels, transistor inverter circuits such as Q2 (low current) or Q5 (high current) may be used for the correct level.

The squelch line should *stay* at the squelch-on level during transmit. This is the case for most transceivers, but if your rig is different, additional squelch gating is required.

time-out circuit

The time-out circuit will shut off the autopatch if it is allowed to transmit continuously for about two minutes. Timer U11 is configured as a missing pulse detector; output is high during transmit. While U11-3 remains high, the input voltage to the Schmitt trigger (U12) will gradually increase to three and a third volts through R35 and C28. U12-3 will go low when this

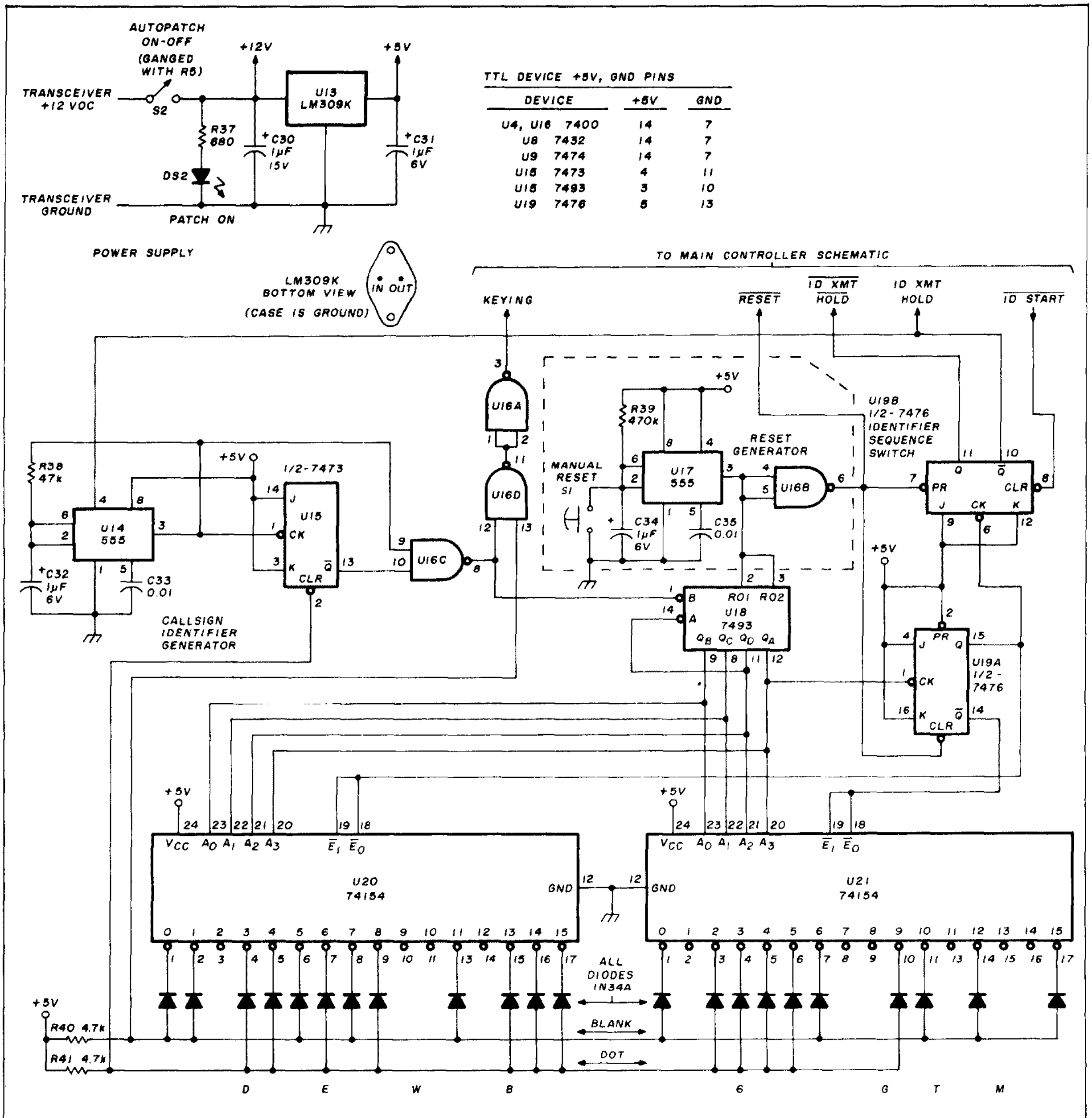

fig. 4. Identifier, reset, and power supply circuits of the autopatch controller. The identifier is programmed for the author's callsign.

voltage is reached, turning the patch off by clearing U9A.

Interrupting the pulses into U11 before two minutes have elapsed will allow C28 to discharge rapidly through CR11 and U11. The time-out cycle is reset. This occurs each time the receiver remains unsquelched longer than the pulse width set by R34 and C26. That width must be longer than the receiver window set by R31. I chose the values of R34 and C26 for any reasonable window width; they should not require adjustment.

You may be wondering why the time-out circuit appears so complex. Why not connect the inverted squelch directly to R35 and C28? The IC-230 and some other transceivers will momentarily unsquelch the receiver when the transmitter turns off during the receive window. This has the effect of periodically discharging C28 and makes the time-out feature use-

less. I designed the time-out circuit to trigger from the window pulses generated by U10 so the circuit will work with all squelch configurations.

Transmit/receive relay K2 connects the telephone line to the transceiver microphone input when picked up. K2 is energized by U11 through Q9. CR9 prevents K2 from being energized when the patch is idle.

identifier

The identifier in **fig.** 4 is a diode-programmable circuit based on a design by K2OAW.[1] Operation has been described adequately in the reference, so I will only cover changes to the original.

The two TTL oscillators of the original were replaced by 555 timers at U14 and at U3 (**fig.** 3), each wired as a minimum-component astable multivibrator. This eliminates start-up problems sometimes experienced with TTL R-C oscillators. R38 controls the ID sending rate and a value of 47K sets the maximum permissible rate of twenty words per minute.

Diode programming of the ID sequence is straightforward. Begin at pin 1 of U20 and connect diodes to the blank or dot lines. No diode and no pin connection produces a dash. DE followed by my call sign is shown as an example.

The 1N34 diodes in both schematics must remain germanium types for a low forward-voltage drop. Silicon diodes have a higher forward voltage and the logic zero level may be too high for proper operation.

U3 in **fig.** 3 generates a 1.5 kHz square wave when the KEYING line is high. Increasing R8 will reduce the frequency. The IC-230 microphone input circuitry filters out the square wave harmonics so the tone sounds like a sine wave. With a transceiver other than the IC-230, check if the ID is audible 15 kHz or more from the carrier center frequency. If audible, add a low-pass filter at the output of U3 or replace the circuit with a sine wave oscillator.

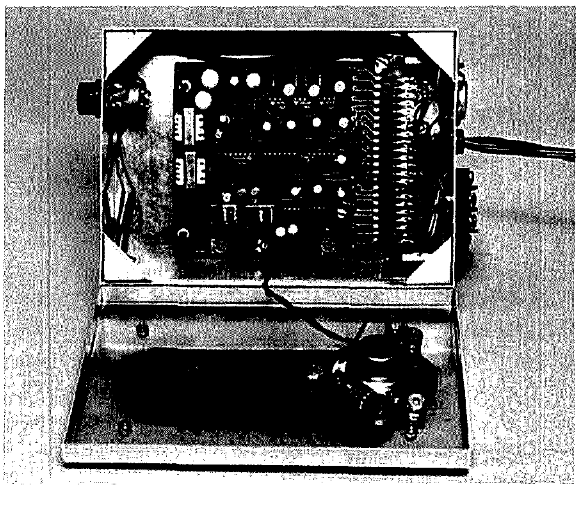

fig. 5. Inside view of the controller chassis. The top circuit board contains almost all of the circuitry of *fig. 3* and mounts by standoffs. Second board is hidden.

reset circuitry and power supply

U17 in **fig. 4** generates a power-on reset for the entire control circuit. The manual reset is a push-button mounted on the controller back panel. The initial state of counter U18 and all flip-flops except the divide-by-two circuit of U15 is reset by U17 and U16B.

Power supply input is in parallel with the transceiver 12 Vdc supply and the autopatch controller adds one-half Ampere to this source. Fuse ratings should be set accordingly. U13 provides regulated +5 Vdc for the TTL and timer circuitry.

construction

The control circuit was built on two separate plug-in boards secured with spacers. I used Radio Shack 276-154 and 276-157 boards with 44-pin edge connectors (276-1551). The monitor speaker is mounted on the detachable bottom plate.

The chassis front panel holds the on-off switch (S2), monitor volume control (R5), in-use indicator DS1, and power indicator DS2.

Fig. 6 shows the chassis rear panel with regulator U13, manual reset push button switch S1, and the telephone interface terminal strip. Transmit and receive audio lines are brought through the grommet and should be shielded cable.

TTL devices may be 74LS types. Two 555 timers may be replaced by a single 556 dual timer with appropriate pin connection changes. All diodes except CR10 and CR7 must be germanium types to avoid exceeding the 0.8 volt logic zero level.

the transceiver

Although my system uses a slightly modified IC-230, other transceivers should work as well if they can shift rapidly between transmit to receive and back to transmit. The rapid change is a requirement of the receive window. This requirement unfortunately excludes many synthesized-frequency units; the synthesizer voltage-controlled oscillator cannot shift transmit and receive frequencies rapidly. The IC-230 is frequency-synthesized but uses fixed-frequency oscillators for the 10.7 MHz first i-f offset.

The IC-230 allows an acceptably-short receive window with no modifications to the solid-state transmit/receive switch. Some autopatch controllers require modification of this switch so that the receiver

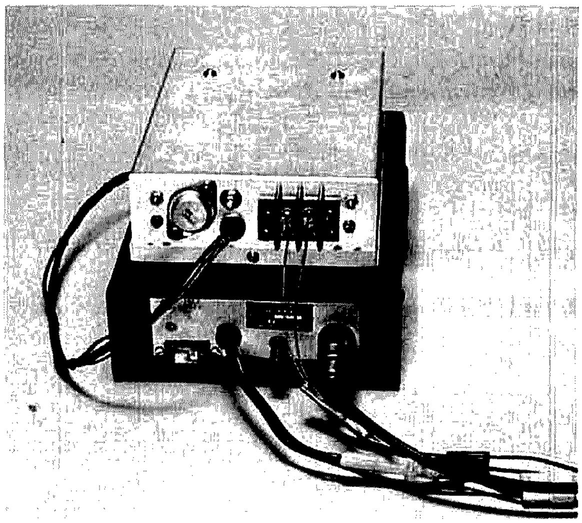

fig. 6. Rear view of the system. The back panel of the chassis contains regulator U13, the manual reset push button, and a terminal strip for the telephone line. The speaker of the IC-230 is disabled by a dummy plug in the external speaker jack to eliminate squelch action noise. The slide switch added to the IC-230 is a modification not needed for the autopatch.

is always on. This makes the rig unsuitable for normal transceiver use: audio feedback occurs during transmit.

One shortcoming of nearly every transceiver is a slow squelch. **Fig. 7(A)** shows the squelch switching circuit in the IC-230, similar to those in other transceivers. Rectified noise enters the base of Q7 to force the collector low, turning off subsequent audio stages and quieting the speaker. Squelch action is slow because C18, C20, and C23 must either charge or discharge each time the squelch changes state. Reducing their values to those in parentheses will speed squelch action considerably.

C21 is an additional low-pass filter element for power to the receiver audio section. Reducing its value will allow the receiver to turn on faster during the controller receive window.

The AF module on the IC-230 chassis bottom contains these capacitors and may be located with the aid of **fig. 7(B)**. In exchange for greater controller complexity, the squelch modification may be avoided altogether by building a separate fast-acting squelch within the controller.

The remaining transceiver modification is that of bringing out the signals needed by the controller. The IC-230 accessory socket provides three of these in addition to ground and +12 Vdc. The squelch switch is the only signal not factory-wired to the socket shown in **fig. 7(C)**. ICOM does provide a connector pin on the AF module, as pointed out in **fig. 8**. A wire from this point will replace the existing accessory socket pin 1 wire and complete the controller interface.

For transceivers without an accessory socket, the microphone input and push-to-talk connections can be made at the microphone connector. Occasionally test pins are provided for discriminator and squelch switch outputs.

primary power and antenna

It may be prudent to use a high-capacity battery in place of the usual line-powered dc supply. It could be on a constant trickle charge and would allow operation during power outages. My system is powered by a 20 Ampere-hour Globe Gel/Cell® and is capable of eight hours continuous transmit after commercial power failure.

In most cases the antenna system should be omnidirectional with as much gain as possible. Remember

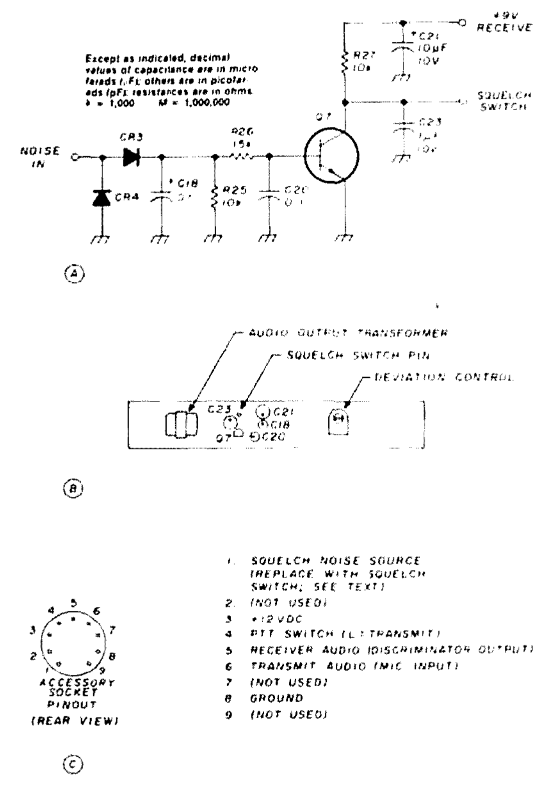

fig. 7. Squelch gating circuit on the AF module in the IC-230 shown in A. Capacitor values in parentheses are changes to shorten the controller's receive window. B shows the locations of components on the AF module. Accessory connector pin connections for the IC-230 are given in C.

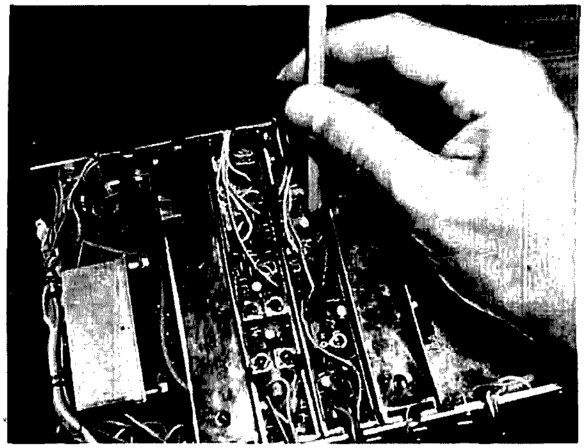

fig. 8. The pencil tip points to the squelch switch pin on the AF module underneath the transceiver.

that the type and location of the antenna is extremely influential toward ultimate coverage area of the autopatch.

control circuit alignment

The only alignment required is the adjustment of nine trimmer potentiometers. R30 and R31 control the receive window, R2 adjusts receiver output into the telephone line, and R4 sets the telephone line level into the transmitter. R18 controls the tone decoder input level while R15, R20, and R24 adjust the individual decoder center frequencies. Decoder frequency stability is dependent on the type of capacitors used for C15, C19, and C22; polystyrene or metal-film types are recommended. R7 sets the ID tone volume.

R2 is set high enough to activate the telephone company's tone-dialing circuitry. R18 is then adjusted for about 100 millivolts audio on pin 3 of each decoder.

An accurate audio generator is handy for aligning the decoders. Lacking this, you can generate the proper tones by the two-button technique on your tone pad. Pressing both * and # produces 941 Hz, * and 7 gives 1209 Hz, while # and 9 will generate 1477 Hz.

operation

Mobile operation is quite convenient. Transmit your call first, then send a # tone. The autopatch will ID as soon as your carrier disappears, then send a dial tone (from the telephone connection) interrupted by the periodic receive window. Wait about a second after initiating your carrier to ensure the system has switched to receive, then send the desired telephone number.

When the conversation is completed (remember the 10-minute limit), send a * to disconnect the patch. Sending a # in place of the * will initiate another call. In either case, the system will ID again when your carrier drops. Should the patch remain on for about two minutes with no received signals, it will time-out and return to standby mode.

Both sides of a conversation may be monitored through the speaker at the autopatch fixed location. Pushing the reset button will return the system to standby. The transceiver may be operated normally by turning off the controller.

Although the system is called a simplex autopatch, it will work as well on an unused repeater channel. This has an advantage of increased privacy since each side of the conversation is on a different frequency.

conclusion

The purpose of this control circuit is to permit a simple, inexpensive, easy-to-construct method for taking advantage of the ability to make telephone calls from mobile or portable transceivers. The system lacks security since anyone who knows or guesses the simple access code can make telephone calls. Three general security methods have been used with autopatches: a complex access code, providing zero and one first-digit lockout, or adding subaudible encoding.[2] Anyone with a tape recorder and some patience can decipher a long access code, so this method is not very secure.

Although requiring more gating and tone decoder circuitry, the zero-one lockout method may be best if the autopatch is to be an open system. Subaudible tone access will provide increased security by limiting patch access.

This autopatch has been in use for several months and I have not experienced problems with it. If you have any suggestions for improving the system, please let me know.

references

1. Peter A. Stark, K2OAW, "A TTL Logic CW ID Generator," *73 Magazine*, February and March, 1973.
2. *The Practical Handbook of Amateur Radio FM and Repeaters*, Bill Pasternak, WA6ITF, with Mike Morris, WA6ILO, TAB Books Inc., 1981.

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ham radio

continued from page 8

on to oppose a no-code license. Several ARRL directors have told me privately that they favor such a license but must oppose it because the members they represent are against it. The same is no doubt true of some ARRL staff members.

Japan, Sweden, and England, among others, have had a no-code license for years and the sky has not fallen. Seventeen other countries require only code "recognition," not reception, for licenses. A number of these licensees go on to learn the code, upgrade their licenses, and use additional modes and frequencies.

The FCC deserves the respect and thanks of thoughtful Amateurs for pursuing this issue in an effort to strengthen the Amateur service and help to ensure its usefulness and its future. — Stuart D. Cowan, W2LX, KM2XDU.

I am astounded, amazed, and dismayed at your September, 1982, editorial position on the code-free license. Your remark that the time has come to discard the "what's good enough for grandpappy is good enough for you" mentality is indicative of your complete lack of understanding of Amateur Radio today and as of yesterday.

Gentlemen, Amateur Radio is not simply another hobby. The creation of a no-code license would be contrary to one of the fundamental traditions of Amateur Radio. Whether or not the prospective licensee plans to operate his station on CW, RTTY, computer, or telephone is unimportant. Learning the code, in the beginning of his interest in Amateur Radio, introduces him to an activity steeped in many traditions; and, to the requirement that such activity be in the public interest, convenience and necessity. The code requirement is the first thing that separates the licensed Radio Amateur from the unlicensed tinkerer and the CB operator, licensed or unlicensed.

Many of us, the writer included, began a professional career in communications via the Amateur Radio route. Even as a teenager we appreciated the importance of our earned entry into the world of Amateur Radio. We became part of a highly special, unique group, with associations of tremendous importance in our chosen career, associations never to be forgotten. — Byron H. Kretzman, W2JTP.

I seriously question your reasons for endorsing a no-code license. We don't need a "larger and faster growing" Amateur population of the quality we have already seen after the recent CB craze. The existing exams are lenient enough, not to mention the multiple choice CW part! To me, anyone who cannot pass a multiple choice CW "test" of 5 WPM is not capable or deserving of being called a ham. When other modes of communication have failed, CW has been used. I would hate to hear about a situation where a ham would not be able to respond to or initiate an emergency message on CW.

The bottom line is your bottom line in your next to last paragraph: "Consider just how such a license could be incorporated into our Amateur structure without cheapening what we've got." — Alan J. Blank, W1BL.

I want to take this opportunity to congratulate you and Joe Schroeder for your joint editorial that appeared in the September, 1982, issue. It was high time that someone told it as it is. I also hold an Extra, have a 30 WPM CP, and still use my old Vibroplex — but, I have a computer and a fairly well equipped small lab, so I don't think I have "given up" trying to stay with the state-of-the-art.

Also, I constantly argue that the FCC test should be slanted more toward R&R, operating procedures, etc., and only the minimum technical questions relating to determining if a transmitter is operating properly. As a graduate E.E., I think some of the questions presently used are slanted toward the hiring exam for an electronics technician, not a ham operator! — John P. Weber, Jr., K4JW.

logic mate

A simple, convenient IC trouble-shooting aid

The **Logic Mate (LM)** is an **IC trouble-shooting aid** that combines the features of a logic monitor and IC tester in a single, compact unit. Inserted between an IC and its circuit socket, it allows pin connection switching, displays pin voltages, and provides output loading. Power is obtained from the main circuit socket and ranges from 5 to 15 volts.

The LM works with all saturating-logic families such as TTL, CMOS, DTL, ECL, plus many linear ICs. U1 pin voltage swings may not be enough with ECL or linear ICs, but all disconnect and forcing inputs operate properly.

about the circuit

Fig. 1 shows the schematic for one of two identical boards in the LM. The plug/header pins connect to the circuit under test while the test socket holds the IC. Pin numbers are indicated for one side of a sixteen-pin dual in-line package (DIP); pin numbers in parentheses indicate the other side.

Two eight-pole, single-throw DIP switches are on the plug/header end. S1 is labelled DISC and allows an individual pin to be disconnected and isolated from the main circuit. S2 is labelled 1-0 and grounds the IC pin under test for a logic 0; resistors R9 through R16 provide a pull-up for logic 1 with any S2 open.

A light-emitting diode (LED) array provides displays of each IC pin under all supply voltage conditions. One DS8654 octal display driver is used for eight LEDs. Individual current limiting for each LED is provided by R1 through R8 and overall current regulation is provided by Q1, CR1, and R17. Without overall regulation, the LED intensity would be too low at 5 volts if the series resistors were selected for 15 volt supplies. Alternately, a correct current at a 5 volt supply would cause too much current at 15 volts and LED life would be reduced.

Individual LEDs need 8 to 12 mA for decent intensity. An LED will drop a relatively constant voltage when conducting. A series resistor selected for 10 mA at a 5 volt supply would cause about a 25 mA LED current at 15 volts.

CR1 is cut off at a 5 volt supply but Q1 conducts via base current through R17 for the LED common cathode return. At higher supply voltages, CR1 conducts and the base of Q1 tracks the increase in voltage. Emitter potential follows base potential for a relatively constant return, or sink current for the LEDs. The emitter return connection provides a low-impedance sink to prevent interaction between different combinations of indicating LEDs.

The octal display driver was designed for a common-cathode display and can source up to 50 mA on each output. The DS8654 will operate in a 4.5 to 33 volt supply range, conservative for this design. Supply voltage is obtained from the main circuit on pin 16 and alternate means are explained later.

construction

Two boards are required. You can etch your own or purchase a pair of etched boards with plated-through holes. If you etch your own, you must solder R9-R16 and U1 pins 1 through 9, and pin 14 on *both* sides of the board. Boards without plated-through holes also require through-board jumpers as indicated in **fig.** 2.

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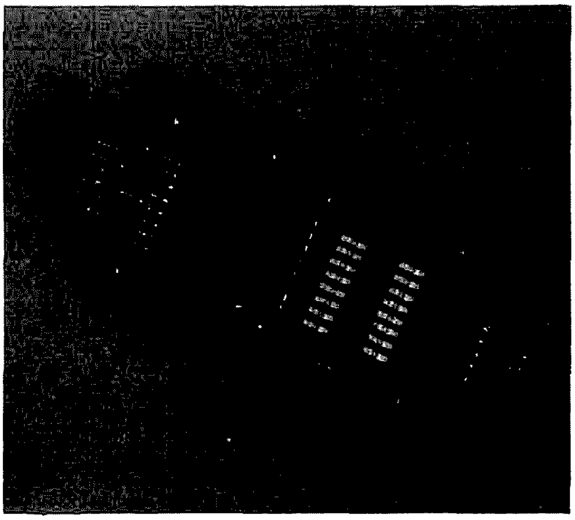
Figs. 2 and 3 indicate R1 to R8 and R9 to R16 as single-inline packages (SIP). SIP resistor packs can be replaced by individual quarter-watt resistors mounted on end. The common connection above the board is made to each SIP end pad; the ninth resistor in each SIP pack is unused and shorted as indicated by the dashed lines in **fig. 1**. Each SIP pack may be mounted in either direction.

The etch pattern and purchased boards allow for three types of LED arrays. The AMP array indicated in the parts list is in DIP form with 0.3-inch pin row spacing. The anode is marked with an A and a dot is at pin 1. The Litronix array has a 0.1-inch row spacing and cathodes are marked with a cross bar. Individual Dialight LEDs mount side-by-side on 0.2-inch spacing. All LEDs must have the anode pins adjacent to U1.

Each board must be jumpered to match the plug/header and test socket pin row. Jumper G must be installed for the board with pins 1 through 8. Ground is obtained from plug pin 8. The board for pins 9 to 16 should have jumper V installed.

Note that jumper V goes directly to S2-8. The reason is two-fold: you cannot accidentally ground the main circuit supply at pin 16, and that pin may be isolated if the supply is not at that pin. The occasional need to ground pin 16 may be accomplished with a clip lead.

Probing points for the main circuit plug/header end are wires soldered on the foil side. I recommend



Top view of one of two completed boards.

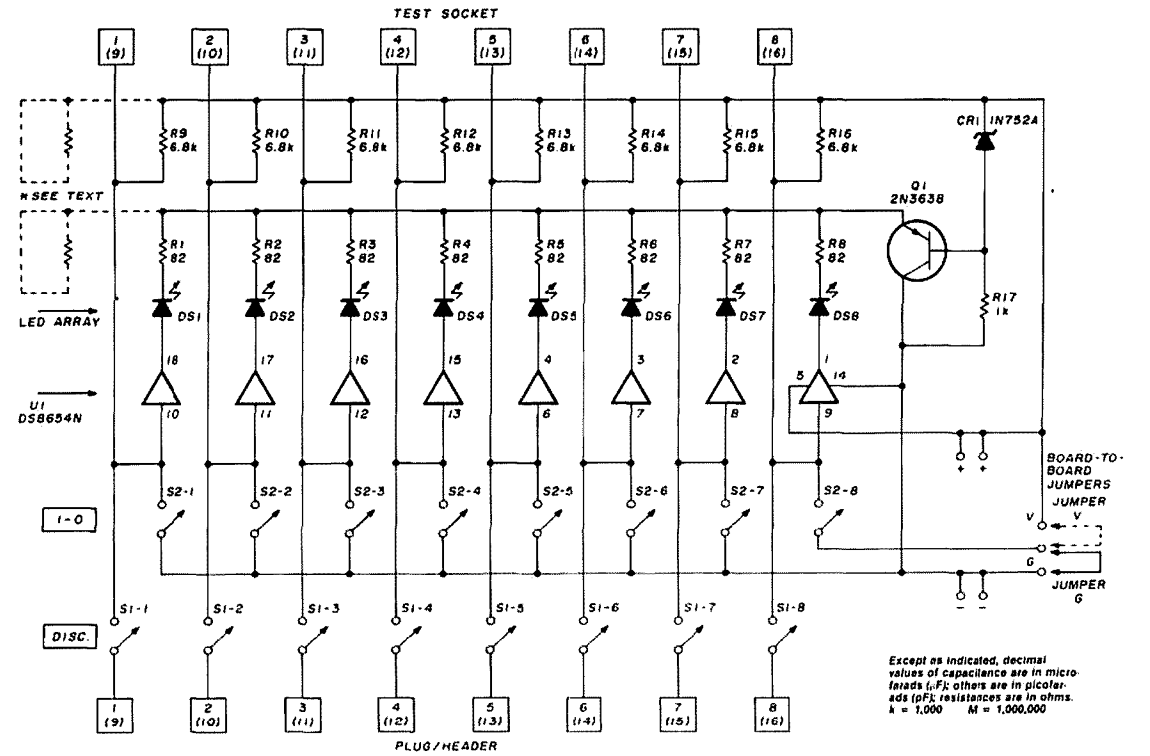


fig. 1. Schematic of one of two identical boards. See text for jumper placement and resistors in dotted lines. LED indicators may be an array or individual units as given in the parts list.

using wire-wrap pins for stiffness and this applies to inter-board jumpers in the final assembly.

final assembly

Both boards are mounted foil-side with both boards inside the header pin rows. I would suggest final assembly in the order following: First, solder two header pins to one board with connecting wires passing through the split in each pin. Take care that header and board connections line up.

Second, insert four board-to-board jumper wires at the pads marked + and − in **fig. 3** and solder them to the first board. Make them perpendicular to the board and long enough to pass through the pads of the second board.

Third, place the second board so the header end is inside the header pins and the board-to-board jumpers pass through the same + and − pads. Recheck alignment and avoid shorts between boards. Solder the board-to-board jumpers when satisfied.

The final step is to solder the test socket to the board. Leave enough socket tail outside the board for probe points. Stiff board-to-board jumpers provide enough support strength for both boards. Additional support is gained with two more jumpers at the unconnected X pads.

A socket on the plug/header will protect the pins between tests. Another option is a zero insertion-force socket for the test socket.

operation

The LM plug/header pin 16 should mate with main circuit socket pin 16 for the positive supply voltage (pin 14 on most 14-pin DIPs). Ground connection is made by closing both S1 and S2 for the ground pin. Pin 8 is the usual ground for a 16-pin DIP, pin 7 for a 14-pin DIP. Remaining pins are controlled by S1 for disconnect or S2 for logic 1 or 0.

For non-standard power and ground connections, a pin ground is always obtained by closing both S1 and S2. The supply pin should have S1 closed and S2 open with the supply voltage clipped to the + board-to-board jumpers. Pin 16 must have its S2 open for a non-standard supply pin; pin 16 may have a main circuit driver and closing that S2 to the supply can destroy that part of the main circuit. Closing S2 of any pin connected to the main circuit supply would yield an unscheduled test of the power supply current limiter!

This description has assumed a positive voltage for the supply with a ground for the return. The LM will also work with negative voltages: consider the LM's + line as the most-positive potential of the main circuit and the − line of the LM as the most-negative. (A negative main circuit supply would require reversal of the V and G jumpers. — Editor's note.)

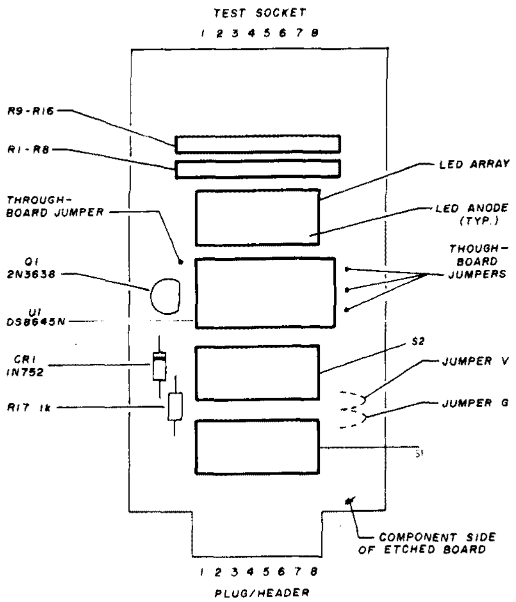


fig. 2. Component placement on each board with foil pattern given in *fig. 3*. See text for jumper placement and use of pads marked +, −, and X.

Parts List for One Board of the Logic Mate

CR1	***1N752A zener diode***
DS1 to DS8	***Choice of:*** ***(1) AMP Inc. 435733-7*** ***(1) Litronix LD468*** ***(8) Dialight 555-3003***
Q1	***2N3638 PNP transistor***
R1 to R8	***82 ohm, 10-pin SIP resistor pack, Bourns 4310R-101-820 or equivalent***
R9 to R16	***6800 ohm, 10-pin SIP resistor pack, Bourns 4310R-101-682 or equivalent***
R17	***1000 ohm, 1/4 Watt, 5 percent resistor***
S1,S2	***Eight-position DIP switch***
U1	***National Semiconductor DS8654N octal display driver***
Test Socket	***Aries A221 or equivalent (wire-wrap tails)***
Plug/Header	***Aries A103 or equivalent***

some examples of operation

One way to isolate a suspect IC pin is to bend it out of contact. This can damage the IC connection even if the IC was good before bending. A better way is to insert the LM and use the appropriate S1 disconnect switch to isolate the pin. The pin is not damaged and the built-in pull-up resistor allows checking of open-collector outputs.

The LM is a logic state monitor if all S1 disconnect switches are closed and all S2 switches opened. The

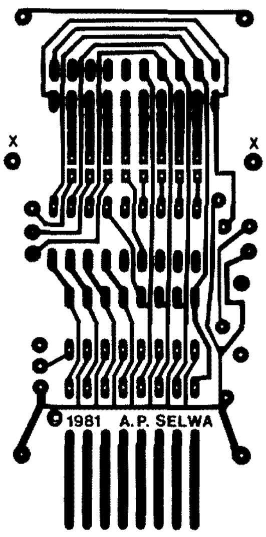


fig. 3. Foil-etch patterns for each double-sided, copper-clad board. Drilled, etched, and plated-through boards are available from PRS Electronics.

IC under test will operate the same as if plugged into the main circuit.

The LM can be an IC tester, isolated from the main circuit. Simply open all S1 switches except for power and ground and plug the LM into any convenient socket. The S2 1-0 switches allow control of all IC inputs.

Linear circuits can be checked by isolating pins with the disconnect switches. Pull-up resistors can provide loads and you can connect clip leads to the test socket probe points. Please note that a linear IC output may or may not indicate the same as a digital device: Voltage swing may be insufficient for the input of driver U1.

where to get parts

Etched, drilled, and plated-through circuit boards are available for $8.00 a pair from:

PRS Electronics
P. O. Box 274
Brownsburg, Indiana 46112.
(Indiana residents add 4% tax.)

The AMP LED array, Litronix LED array and the DS8654 are available COD from:

Advent Electronics, Inc.
Dept. 7
8446 Moiler Road
Indianapolis, Indiana 46268.

ham radio

forget memory

Ni-Cd memory myths discharged and facts unveiled

Ni-Cd memory myths rank with SWR stories and VHF intermod misunderstandings as the most common misconceptions in Amateur Radio. I've had hams tell me they know all Ni-Cds eventually get a memory problem because "it says so in the GE battery book." Not true!

The half-inch thick General Electric Ni-Cd handbook[1] devotes only three paragraphs to the topic of memory effect. Two paragraphs state and restate that "the memory effect does not manifest itself when the battery is discharged to random depths of discharge or overcharged for random amounts of overcharge time as is typically the case in most applications." That statement should relieve hams, not alarm them.

If your Ni-Cd cell voltage falls to zero quickly when discharging, the problem isn't memory effect. A memorized cell experiences voltage drops early in its discharge cycle, but instead of dropping to zero, it goes to the usual discharge cutoff point of around 1.0 volt. For the remainder of the discharge cycle, the cell still provides power but at an unacceptably low voltage.

Two General Electric engineers presented a scholarly treatise on the memory effect in the IEEE *Spectrum* magazine about five years ago.[2] They told how the effect was discovered in satellite test programs, where the cells received a carefully controlled charge/discharge regimen at a constant temperature for hundreds of cycles. Even then, the effect did not always appear in all types of cells. They concluded that "though sintered plate nickel-cadmium batteries can remember, the conditions necessary are almost *never* encountered in practice" (emphasis added). Sintered plate batteries are the kind we hams use.

So, if that isn't the problem, why didn't your last Ni-Cd pack last for a thousand charges? There are a couple of likely causes — cell reversal and sustained overcharge.

Do you stop talking the minute your HT's low battery indicator comes on? If you don't, you may be sending your pack to an early grave. Why? Not all cells in a Ni-Cd ppack are identical, nor do they reach complete discharge simultaneously. If one cell goes to full discharge first and you draw a medium-to-heavy load, the good cells attempt to charge the discharged cell in the reverse direction, which can damage it quickly, usually causing it to short. A short can be zapped out of a Ni-Cd, but full cell capacity is seldom restored. The zapped cell is likely to get reverse-charged again the next time the pack charge gets low.

Don't continue to use your rig once the low battery warning is present, or you may kill one or more cells through reverse charge. Also, if you've been trying to prevent the memory effect by frequently discharging the entire pack to zero with a single load, you can see that you're actually doing more harm than good.*

Now the other problem: can you really overcharge a battery with that little wall charger? Sure! Next time you take your battery off the charger, feel the pack (not the charger). Isn't it warm? Try it for two or three days! (On second thought, don't!) When the battery reaches full charge, and you keep putting current into it, the internal chemical reaction changes. The current from the charger begins to break down the water in the cell into hydrogen and oxygen. Much of it recombines, but even at typical slow-charge rates (about 0.1C), heat and pressure build up. When this happens, the cell's vent opens and it slowly dries out. If you're a typical HT user who uses only a fraction of the battery capacity each day, but charges it for 15 hours each night, you shouldn't be surprised if your batteries require replacing prematurely.

Extended slow-charging also can produce what the General Electric engineers call a sustained overcharge effect.[3] It's a depression of the discharge voltage at some point after a long overcharge, and is quite common. The longer the overcharge, the ear-

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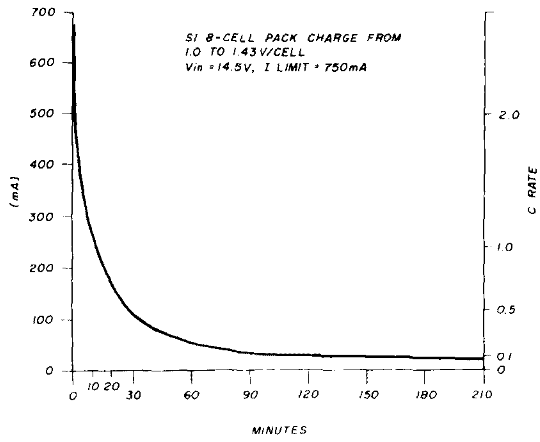


fig. 1. Charge current versus time for the taper charger described in the June, 1981, *QST*, by KØOV. Note the slow-charge current of 25 mA is reached at two hours, and the current continues to decrease, protecting the battery. This data was taken on a 250 mA-H pack used in early models. Rigs that use 450 mA-H packs will take longer to charge at this rate, but the maximum rate can be increased to compensate.

lier in the discharge cycle the step-down in voltage will occur. To the user it appears just like the memory effect described earlier, but is from a different cause. If your battery pack has shown symptoms of memory, chances are it's because it was overcharged. The traditional memory cure also works for the sustained overcharge effect.

What is needed is a way to stop charging the pack when a full charge is reached. W2GZD described such a slow charger in the October, 1981, *QST*.[4] When a specific voltage (1.43 volts per cell) is reached, the charger automatically cuts back to a trickle charge mode that doesn't cause heating.* That's nice for a radio that sits on a charger 24 hours a day, but what about the ham-on-the-go who needs a quick charge? A good, safe, fast-charge method was described in the June, 1981, *QST*.[5] The unit charges a pack fully in only a couple of hours, and has a meter that shows the charging progress (**fig. 1**). Even though this is a fast charger, the battery doesn't warm up because the current tapers down to a slow charge value automatically as full charge is approached. The radio can remain in use during the fast charge. With a unit like this, you get the best of everything, fast charge, long battery life, and convenience.*

One final caution about your charging system — beware of extreme temperature environments. At temperatures below about +5 degrees Celsius, overcharge quickly results in a buildup of hydrogen pressure which can result in cell damage. At very high temperatures, charging is not very efficient and cell temperature can rise to damaging levels during fast charging. In both extremes, the end point for charging is no longer 1.43 volts, meaning that a taper charger may not properly sense full charge. For best results, avoid trouble by doing your charging with the cells at or near room temperature. Don't leave your HT in a freezing car all night and charge it without allowing the cells to warm up first.

Here's how to make your Ni-Cds last:

1. Don't deliberately discharge packs to zero.
2. Don't continue to use your HT after the low battery indicator comes on.
3. Don't use a wall charger for more than 15 hours per charge.
4. Don't do a 15-hour slow charge on a battery that is only slightly discharged.
5. Do monitor pack voltage when slow-charging a partially discharged pack. Stop charging when voltage reaches 1.43 volts per cell, measured at room temperature with 0.1C charging current applied.
6. Don't charge your pack in freezing temperatures or in direct sunlight.
7. Don't continue to charge a battery if it gets warm during charging. Check for a shorted cell or misadjusted charger.
8. Do build or buy a taper charger for fast charging and operating with convenience and safety.

You won't need to worry about memory if you follow this advice.

references

1. General Electric Company, *Nickel-Cadmium Battery Application Engineering Handbook*, Second Edition, 1975, pages 7-8.
2. S.F. Pensabene and J.W. Gould, "Unwanted Memory Spooks Nickel-Cadmium Cells," *IEEE Spectrum*, September, 1976, pages 32-36.
3. General Electric Company, pages 6-9 and 6-12.
4. D.W. Potter, "Those NiCad Batteries and How to Charge Them," *QST*, October, 1981, pages 34-35.
5. J.D. Moell, "A Fast Charger and Regulator for the Tempo S-1," *QST*, June, 1981, pages 34-35.

ham radio

*Discharging the cells individually is recommended by some authors as a memory cure that won't cause cell reversal, but you should make sure the battery has the disease before applying the cure. Discharging cells deliberately in hopes of preventing memory is a complete waste of time and effort. This is the traditional memory cure.

*Many hams erroneously call their little wall charger a trickle charger. Actually, the wall charger performs a slow charge (about 0.1C). A true trickle charge (0.02C or less) is used only to keep a battery charged. See Chapter 5 of the General Electric book for more details on rates of charge.

*Don't confuse this circuit with regulators or battery beaters, which only operate the rig and don't charge the battery.

ham radio TECHNIQUES

Bill Orr W6SAI

Long, long ago, in a galaxy far, far away, or so it now seems, the World Administrative Radio Conference set aside certain new, narrow, high-frequency bands for use by the Amateur Service. Since January a year ago, over sixty countries have permitted Amateur operation in one of these bands, 10.1 MHz to 10.15 MHz. But the United States — whose Amateurs spearheaded the effort at WARC to get the new bands — has dragged its feet on granting permission to operate in these bands. The reasons for the long delay make an interesting story, indeed.

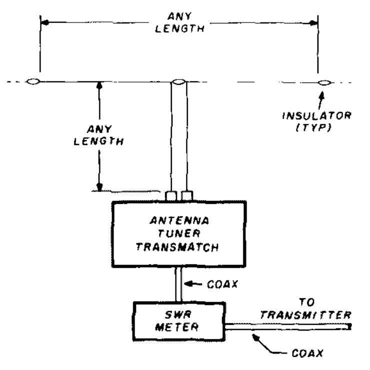


fig. 1. A center-fed wire serves as an all-band antenna when used with a suitable antenna tuner located at the station. Lengths of antenna and two-wire transmission line are not critical as long as overall length of wire is at least one-quarter wavelength at lowest frequency used.

Finally the combined efforts of the ARRL and a hot letter to the FCC from Barry Goldwater, K7UGA, opened the door and (as of this writing in September) it looks as if the 10-MHz band will be opened to U.S. Amateurs around the first of 1983, if not before. And that's good news for 1983!*

The 10-MHz band is full of interesting DX when conditions are good, and one of the first questions raised by prospective 10-MHz operators concerns antennas for the new band, particularly all-band antennas that will cover existing bands plus the new ones. That means coverage of the 160, 80, 40, 30, 20, 17, 15, 12, and 10 meter bands, now available on the bandswitch of many of the new transceivers!

simple all-band antennas

The first all-band antenna that comes to mind is the well-known center-fed long wire (**fig. 1**). Used with an open-wire transmission line and an antenna tuner at the station, this simple antenna will work well on any frequency within the range covered by the tuner. (The tuner is sometimes called a Transmatch.) The tuner is coupled to the station equipment via a coaxial line and SWR indicator.

A second simple wire antenna that will cover all the Amateur high-frequency bands is the end-fed wire (**fig. 2**). A pi-network composed of a rotary inductor and two capacitors matches the wire antenna to a 50-ohm system.

Users of the end-fed antenna know that under certain conditions the antenna will tune up well, but the shack will be full of rf and feedback. This can cause erratic operation of the equipment, TVI, and other unpleasant problems. The cause of the diffi-

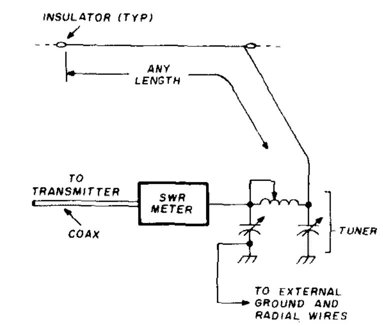


fig. 2. End-fed, random-length wire with simple pi-network tuner serves as all-band antenna. A good ground system is required with this antenna.

*As of 3 pm EDST, October 28, U.S. Amateurs were permitted use of 10.100-10.109 and 10.115-10.150 MHz using A_1, F_1 modes at a maximum power level of 250 watts.

culties is that the equipment is not at rf ground potential. Attaching a ground to the equipment usually doesn't help a bit, as the inductance of the ground wire upsets the situation even more. The use of quarter-wavelength radial ground wires cut to the operating frequency will solve this vexing problem. The radial ground wire is merely a length of insulated wire, free at the far end. It is cut to an electrical quarter wave at the operating frequency and affixed to the ground post of the equipment. The wire can be tossed on the floor behind the operating desk.

the Australian broadband dipole

One of the best all-band antennas in use is the so-called Australian dipole, which I briefly mentioned in an article in *CQ* magazine, October, 1974. After this article, the antenna sank into oblivion, at least in the United States.

In spite of this seeming lack of interest, the unusual antenna has continued to be used by Amateurs and commercial point-to-point services in other areas of the world. It eventually caught the attention of D.W. Harris (A22BX), the Deputy Director of Broadcasting (Engineering) at Radio Botswana in Southern Africa.

In common with many developing countries, Botswana has internal communications difficulties. Roads are often poor in rural areas and the telecommunications networks are hardly developed. As a result, the news service of Radio Botswana is an important facet in passing information from remote districts to the capital.

Harris decided to use high-frequency SSB transceivers for this purpose — along with broadband antennas — with provisions for patching tape recorders into the transceivers. (Drake TR-7 transceivers were used with the SL-4000, 4-kHz passband filter.)

The problem of a broadband antenna which could be easily built and installed was formidable, as it had to provide less than a 2:1 SWR across the operating range. An article on the "Broadband Travelling Wave Dipole" appeared in the April, 1974, issue of *Amateur Radio* (Australia), which described an interesting antenna developed for use in the Australian Outback for the Flying Doctor radio service.

Harris and his staff built several Australian dipoles and tried them out with varying success. Finally, a modification of the original design produced a noncritical, wideband antenna which, with a special balun, showed less than a 2:1 SWR from *below* 3 MHz to over 20 MHz. The antenna was useful up to 30 MHz, as shown in the SWR plot of **fig. 3.**

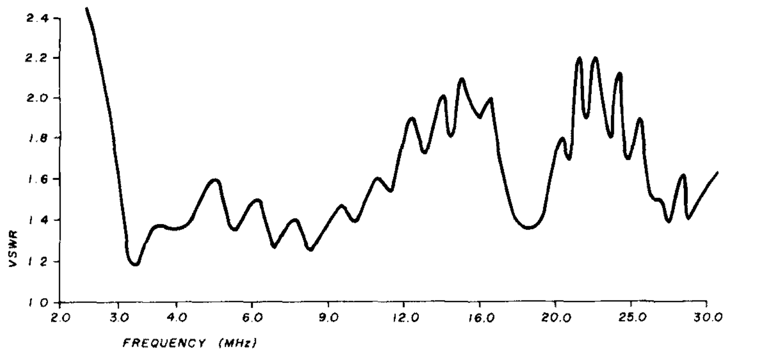


fig. 3. Measured performance of Australian Dipole, showing VSWR plotted against frequency every 0.5 MHz from 6 MHz to 30 MHz.

the modified Australian dipole

A simplified diagram of the A22BX version of the Australian dipole is

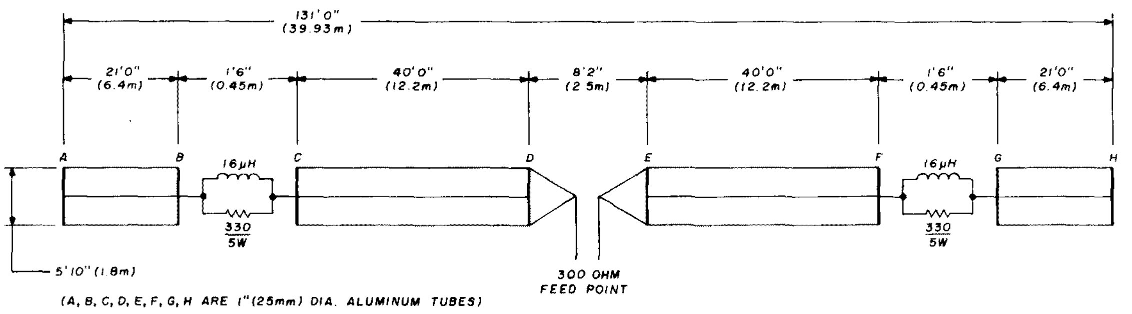


fig. 4. Three wire "flat-top" with loading networks provides broadband service between 2 MHz and 30 MHz. Antenna is fed with balun and coaxial line.

shown in **fig. 4**. Overall length is about 133 feet (40 meters). The antenna consists of a flat-top of three parallel wires, broken at intervals by simple loading networks placed in series with the wires. The feedpoint impedance of the antenna is about 300 ohms. A toroid transformer wound on an Amidon two-inch diameter red core was used to make the impedance transformation to a 50-ohm line (**fig. 5**).

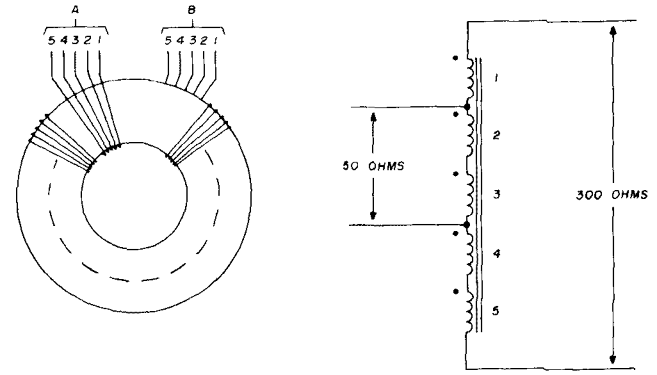


fig. 5. Transformer is wound on iron-powder toroid. Fourteen turns, each composed of five parallel wires, are wound on the core. Turns are interconnected as shown at right to provide 5/2 turns ratio (6.25/1 impedance ratio).

Details of antenna construction are shown in **fig. 6**. Hard-drawn copper wire is used for the flat-top to prevent stretching. The small networks are made of a 300-ohm, 5-watt composition resistor placed in parallel with a small inductor.

To hold the three wires of the antenna in position, yet allow easy handling on the ground, the networks are fastened to a framework that is big enough to attach the antenna wires to, as shown in the detail drawings. One-inch diameter PVC plastic conduit is used for the insulators. The aluminum spreaders were made from decorative aluminum L-shaped stock measuring about an inch wide and ½-inch thick, used for edging on Formica kitchen table tops. Small, 10-32 machine bolts, nuts, and hardware are used to hold the various strips and tubes together.

A22BX suggests using polypropy-

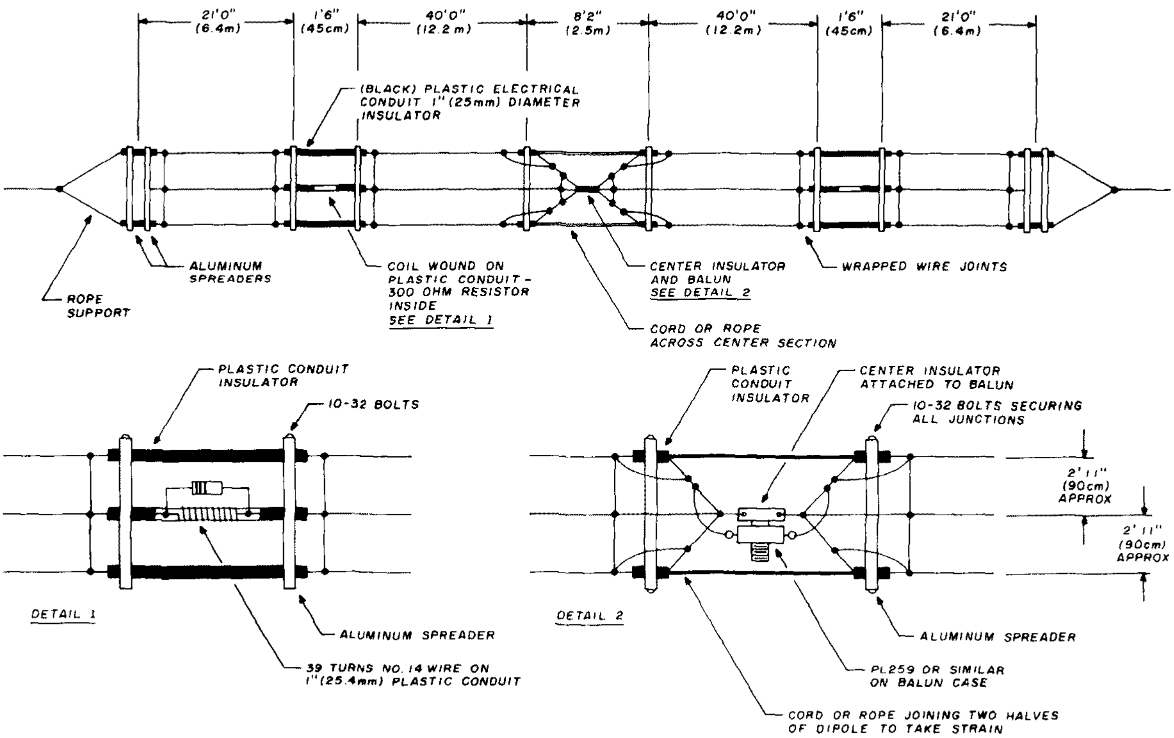


fig. 6. Assembly details for Australian Dipole.

lene plastic fiber rope in the antenna assembly to resist the effects of ultraviolet light. A suitable substitute rope is UV-resistant polyester material.

antenna installation

The SWR curve shown for this antenna was measured through about 75 feet (23 meters) of coaxial line, and the antenna was suspended in the air at about 40 feet (12 meters). As can be seen, the SWR was excellent up to 15 MHz, rising between 15 and 16 MHz to 2.1:1, then dropping down to low values up to 21.5 MHz, where two SWR peaks at 2.2:1 appear. The SWR curve then gradually drops off to a low value at 30 MHz. Undoubtedly the antenna is also suitable for operation above 30 MHz, but higher frequency measurements were not made.

The SWR can be adjusted around 16 and 22 MHz by varying the height of the antenna above ground, and for a permanent installation, the ends of the Australian dipole can be varied in height to smooth out the SWR curve by taking advantage of ground reflection.

the balun transformer

The matching transformer is wound on an iron-powder core having an outside diameter of two inches (5.08 cm). Inner diameter is 1¼ inches (3.18 cm). It is made by Micrometals Corp. and has the Amidon part number T-200-2. It is coded red, and has a permeability of ten. It is recommended for operation over the range of 1 MHz to 30 MHz. The core is also sold by J.W. Miller Division of Bell Industries as part number T200-2.

To prepare the core, wrap it with a layer of 3M brand (or equivalent) glass epoxy tape to prevent arcing between winding and core. A single winding composed of five parallel wires is placed on the core. No. 14 AWG Formvar-insulated (about 1-mm diameter) wire is used. Fourteen turns of the five-wire combination are wound on the core. The approximate length of wire used for each winding is about 5 feet (1.5 meters).

It is easier to wind the core than to explain how it's done. One set of wire ends is held in a vise and the five wires are smoothed out until they lie parallel. The parallel group of wires is stretched to remove kinks and then removed from the vise. The wires can be wound on the core all at once, or three wires can be wound on, followed by two, if that seems more convenient. In either case, the windings should all lie together.

When completed, continuity of each winding can be checked with an ohmmeter and the wire ends marked for convenience with a drop of epoxy paint. The last step is to interconnect the windings to get the proper transformation ratio. The windings are connected in series and the 300-ohm termination taken from the ends of the windings. The 50-ohm input points are tapped off between the ends of the second and third windings. This provides a turns ratio of 5:2 and an impedance transformation of 6.25:1.

When the transformer is completed it is given a coat of casting resin to protect it from the weather.

The transformer is attached to the center insulator of the antenna and a coaxial receptacle (SO-239), or a waterproof type-N connector, affixed to the balun terminals, and mounted to the center insulator.

When the antenna is completed, it should be raised in position and adjusted to provide the lowest value of SWR in the most important frequency regions of operation.

Note: This antenna is based upon a design by Dr. R.J.F. Guertier and G.E. Collyer of Antenna Engineering Australia (Pty.), Ltd. and was described in *Amateur Radio* (Australia), April, 1974. Information on the Australian Dipole is gathered from issues of *Amateur Radio*, the monthly publication of the Wireless Institute of Australia, Box 150, Toorak, Victoria 3142, Australia.

ham radio

the ham notebook

switching circuit

I own a Chevy Vega in which I installed an ICOM IC-255A. I quickly discovered the tiny internal speaker, which was aimed at the floor of the car, could not compete with car and road noise. I couldn't afford to reinsulate the car and the Hatchback design wouldn't support an external speaker. However, there was already one well-placed speaker in the car — the one used for the built-in a-m radio.

I could have installed a switch easily, but that would have generated another problem: 2 meters in the Las Vegas area is not teeming with activity. I only needed to use my car's built-in speaker when 2 meters was active. I needed to be alerted when my ICOM was receiving so I could switch the car's speaker from the a-m radio to the ICOM. Going one step further, why not find some way to automatically switch between receivers?

The IC-255A has an accessory socket that supplies 7 volts when the squelch is on. I tried using a transistor as a switch controlling a relay with my squelch voltage controlling the transistor. This was unsuccessful, it played havoc on my audio output from the ICOM. Dale Porray, AD7K, a local ham, came up with the perfect solution, using a voltage comparator circuit. The final circuit is shown in **fig. 1**.

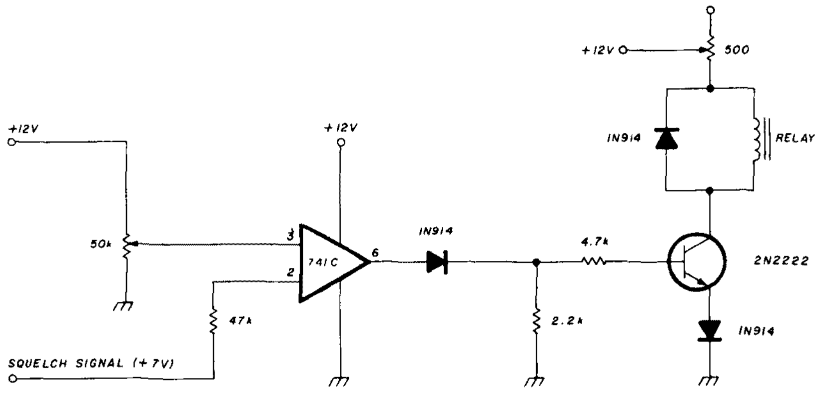


fig. 1. Voltage comparator circuit.

The 741C monitors the difference between the squelch signal and the reference voltage obtained from the 50K pot. It also provides good isolation for the ICOM squelch circuit. Once a signal breaks the squelch, the 7 volts disappears and the 741C causes the 2N2222 to deactivate the control relay. The 500 ohm pot on the relay's hot side is adjusted to prevent the relay from latching up. By obtaining 12 volts from the automobile and *not* the IC-255A accessory socket, the circuit activates only when the car's ignition is on. Since my IC-255A is wired directly to the battery to supply continual voltage to the memory circuits and to provide operation when I'm not driving, I used the relay configuration shown in **fig. 2**. The relay I used is Radio Shack part # 275-206. The 10 ohm resistors are 2 watts.

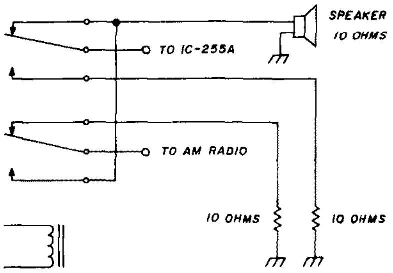


fig. 2. Continual voltage relays.

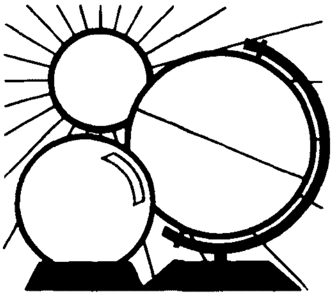
When the car's ignition is off, the IC-255A is wired directly to the car's speaker via the relay, and no unnecessary battery drain is made by operating the relay. When the car's ignition is on, I can enjoy listening to my car's a-m radio, knowing I will hear any signals coming through my IC-255A because my a-m radio will be muted.

This circuit can also be used to mute one radio in favor of another.

Construction and layout of the circuits are not critical. I used perfboard and direct wiring. I would recommend using an IC socket for the 741C. I got fancy when I wired the unit into the car: by using plugs and snap-on connectors, the entire unit can be removed and the speaker reconnected to the a-m radio in a matter of minutes.

This circuit is not unique to the IC-255A and can be applied to any receiver where a voltage is present when the squelch is on.

Fred Dahnke, WB6IQV



DX FORECASTER

Garth Stonehocker, K0RYW

last minute forecast

The higher frequencies (10-20 meter bands) are expected to provide excellent DX openings during the third week of this month. Openings will probably build up only the first two weeks and fall off the last week of the month. The lower frequency bands (nighttime, 40 to 160 meters) will probably have very good DX throughout the month and enhanced DX the first and last week. Disturbed conditions may be enhancing trans-equatorial paths and creating east/west path problems around the 5th, 14th, 24th, and 31st of the month.

Lunar perigee and full moon are on January 28 this month. There will be an intense but short meteor shower lasting a few hours some time between January 2nd and 4th. It is known as the Quadrantid shower.

Many of the DXers using the radio propagation quality formula (August, 1982, DX Forecaster) have expressed an interest in how the formula came to be. I thought I'd pass this along:

The North Atlantic Radio Warning Service (NARWS) of the National Bureau of Standards of the Department of Commerce provided this service via WWV broadcasts from World War II until 1978. The forecast was done for every six hours; that is, four times daily. The last two hours of monitoring of each six-hour period was the basis for the next six-hour forecast. Radio circuits crossing the North Atlantic were monitored and evaluated by the FCC, RCA, ITT, the Navy, the Coast Guard, the Canadian Broadcasting Corp., and NARWS. The forecast was also based on solar-geophysical and ionospheric conditions which were called and monitored by the NARWS.

Statistical analysis was done to keep the evaluations within standard-deviation limits. Mr. Harris, one of the forecasters, noticed over the years that two of the useful predictors were the radio flux values and geomagnetic A or K variability indices. He made computerized comparisons of flux values and variability indices to the observed radio quality monitored evaluations from 1947 (the start of Ottawa, Canada, 2800-MHz radio flux measurements). The mathematical formulation was obtained from these correlations. The calculated quality values from the formula were then compared with the observed quality values over the span of years from 1947 into the latter 1970s; they matched well in summer but toward each winter the calculated quality was greater than the observed. Therefore utilizing a modeling technique, the seasonal term, θ, was developed, using the day of the year and a cosine function to gradually lower the calculated number. An earlier version of θ was inadvertently given in August. The correct formula is: $\theta = 1.0 - 0.2625 \cos^2 0.49315X$.

An attempt was made to determine why θ was needed, but several effects could have been involved: enhanced sporadic-E signal strength in summer; lowered F2 region heights (therefore more hops required across the Atlantic) in the high latitude trough in winter; or higher winter absorption (known as the winter absorption anomaly) than the closer-sun higher radio flux can account for. These are postulations that have been proposed but not fully researched since it is very difficult to sort out individual geophysical mechanisms operating in such cases.

band-by-band summary

Ten meters will be open occasionally for F2 long skip by the trans-equatorial one-long-hop propagation mode (TEM). The openings will follow the sun during the day and into late evening. Geomagnetic disturbances will enhance this mode, as will high solar flux. Openings may favor southern Africa, South America, and Australia — particularly southern Africa.

Fifteen Meters can have the same TEM modes as 10 meters. The openings should be more frequent. Worldwide DX is prevalent from after sunrise until well after sunset during the periods of high solar flux (listen to WWV at 18 minutes after the hour for reports on solar and geomagnetic conditions). A good practice when bands are open is to work the highest band that is open first, then drop down in frequency to catch each band until it closes.

Twenty meters will be open most days and nearly through the night to some areas of the globe, with long skips of 1000-2500 miles and plenty of short-skip of 1200 miles near midday. Both propagation modes follow the sun across the sky: east, south, then west. This is the workhorse of the bands for DX as well as traffic handling.

JANUARY

WESTERN USA

GMT	PST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	4:00	10	20	15	15	20	10	10	10
0100	5:00	15*	20	15	15	20	10	10	10
0200	6:00	15	20	15	15	15	15	10	15
0300	7:00	15	40	20	15	15	15	15	15
0400	8:00	20	40	20	20	15	15	15	20
0500	9:00	—	40	20	20	15	20	15	20
0600	10:00	—	40	—	20	15	20	20	40
0700	11:00	—	—	—	20	20	20	20	40
0800	12:00	—	—	20	20	20	20	20	40
0900	1:00	—	—	20	40	20	20	20	40
1000	2:00	—	—	—	40	40	40	—	40
1100	3:00	—	—	—	—	40	40	—	—
1200	4:00	—	—	—	—	40	40	40	—
1300	5:00	—	—	20	—	—	40	40	—
1400	6:00	—	20	20	—	—	40	—	—
1500	7:00	—	20*	15	15	—	—	—	—
1600	8:00	20	15	10	10	—	20	20	—
1700	9:00	20	15	10	10	—	—	20	—
1800	10:00	20	15	10	10	—	15	15	—
1900	11:00	15	20	10	10	20	10	15	—
2000	12:00	15	20	15	10	15	10	10	20
2100	1:00	15	20	15	10	15	10	10	20
2200	2:00	15	40*	15	15	15	10	10	15
2300	3:00	10	20	15	15	15	10	10	15*

MID USA

GMT	MST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CST
0000	5:00	15	20	15	15	20*	10	10	10	6:00
0100	6:00	15	20	15	15	15	10	10	15	7:00
0200	7:00	15	40	20	15	15	10	15	15	8:00
0300	8:00	20	40	20	15	15	15	15	20	9:00
0400	9:00	—	40	20	20	15	15	20	20	10:00
0500	10:00	—	40	20	20	15	20	20	40	11:00
0600	11:00	—	40	20	20	20	20	20	40	12:00
0700	12:00	—	40	20	20	20	20	20	40	1:00
0800	1:00	—	20	20	20	20	40	40	40	2:00
0900	2:00	—	20	—	40*	20	40	40	40	3:00
1000	3:00	—	—	—	40*	20	40	—	—	4:00
1100	4:00	—	—	—	—	40	—	—	—	5:00
1200	5:00	—	—	—	—	—	—	—	—	6:00
1300	6:00	—	20	20	15	—	—	—	—	7:00
1400	7:00	20	15	15	10	—	—	—	—	8:00
1500	8:00	20	15	10	10	—	—	—	—	9:00
1600	9:00	20	15	10	15*	—	—	20	—	10:00
1700	10:00	20	15	10	15	—	20	20*	—	11:00
1800	11:00	20	15	10	10	—	15	15	—	12:00
1900	12:00	20	20	10	10	—	15	15	—	1:00
2000	1:00	—	20	10	10	20	10	10	20	2:00
2100	2:00	—	20	10	10	20	10	10	20	3:00
2200	3:00	—	40*	15	15*	20	10	10	15	4:00
2300	4:00	—	20	15	15	20	15*	10	15	5:00

EASTERN USA

GMT	EST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	7:00	15	20	15	15	15	15	10	15
0100	8:00	15	20	15	15	15	15	15	15
0200	9:00	15	20	20	20	20	15	15	20
0300	10:00	20	40	20	20	20	15	20	20
0400	11:00	—	40	20	20	20	20	20	20
0500	12:00	—	40	20	20	20	20	20	40
0600	1:00	—	40	20	20	20	20	40	40
0700	2:00	—	40	—	20	20	20	40	40
0800	3:00	—	40	—	40	20	40	—	20
0900	4:00	—	20	—	40	—	—	—	20
1000	5:00	—	—	—	40	—	—	—	—
1100	6:00	—	—	20	40	—	—	—	—
1200	7:00	—	—	15	40	—	—	—	—
1300	8:00	20	—	10	15	—	—	—	—
1400	9:00	20	10	10	15	—	—	—	—
1500	10:00	20	10	10	15	—	—	20	—
1600	11:00	15	10	10	15*	—	—	15	—
1700	12:00	15	15*	10	15*	—	20	15	—
1800	1:00	—	15	10	15*	—	15	15	—
1900	2:00	—	15	10	10	20	15	15	—
2000	3:00	—	20	10	10	20	10	15*	—
2100	4:00	—	20	10	10	20	10	10	20
2200	5:00	—	40*	15	10	20	10	10	15
2300	6:00	15	20	15	15	20	10	10	15

*Look at next **higher** band for possible openings.

Forty meters is the transition band into all-night propagation as well as some short skip during the day. Most areas of the world can be worked from darkness till just before sunrise. Hops shorten on this band to about 2000 miles, but the number of hops can increase since signal absorption is low during the night.

Eighty meters is traditionally a rag-chewer's band but much DX work is also possible. The band operates much like 40 meters except that the hop distances shorten to about 1500 miles. Noise from distant thunderstorms is so low as to make these bands a joy to work this time of year. The path direction follows the darkness across the earth (east, south, then west). Just wiggle in between the QRM.

One-sixty meters will be about like 80 meters, with reduced range to 1000 miles. It provides good DX for enthusiastic DXers. The new band power and frequencies should increase activity here, so we'll be listening.

ham radio

MOBILE IGNITION SHIELDING. Estes Engineering, 930 Marine Dr., Port Angeles, WA 98362

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. PO Box RY, Cardiff, CA 92007

WE WANT your innovative, working antenna and ham product designs for development and marketing. Jon Titus (KA4OVK) Box 242, Blacksburg, VA 24060. (703) 951-9030.

WANTED: Buy or swap Model 28 paper winders (LPW 300). Top prices paid for complete units or parts. Call or write Van, W2DLT. In NJ 800-272-1331, outside NJ 800-526-3662. Box 217, Berkeley Hts., NJ 07922.

MANUALS for most ham gear made 1937/1970. Send $1.00 for 18 page "Manual List", postpaid. HI-MANUALS, Box R802, Council Bluffs, Iowa 51502.

SATELLITE TELEVISION — Howard/Coleman boards to build your own receiver. For more information write: Robert Coleman, Rt. 3, Box 58-AHR, Travelers Rest, SC 29690.

WANTED: Schematics-Rider, Sams or other early publications. Scaramella, P.O. Box 1, Woonsocket, RI 02895-0001.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

TELETYPE PARTS WANTED: Any quantity, any models, highest prices paid. In NJ 800-272-1331, outside NJ 800-526-3662. Van, W2DLT, Box 217, Berkeley Hts., NJ 07922.

SATELLITE TELEVISION: Discount prices on all major TVRO items. Communications Consultants. (501) 452-3149.

RUBBER STAMPS: 3 lines $3.25 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

WANTED: New or used MS and coaxial connectors, synchros, tubes, components, military surplus equipment. Bill Williams, PO #7057, Norfolk, VA 23509.

VERY In-ter-est-ing! Next 5 issues $2. Ham Trader "Yellow Sheets", POB356, Wheaton, IL 60187.

CB TO 10 METER PROFESSIONALS: Your rig or buy ours — AM/FM/SSB/CW. Certified Communications, 4138 So. Ferris, Fremont, Michigan 49412; (616) 924-4561.

HAMS FOR CHRIST — Reach other Hams with a Gospel Tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

KT5B Multi-Band dipole 160-80M (WARC) $59.95, 2KW plus center connector w/SO239 $8.50. Kilo-Tec, P.O. Box 1001, Oak View, CA 93022. (805) 646-9645.

MOTOROLA ALL SOLID-STATE MOTRAN RADIOS. Model X43LSN-2170, four frequency, transmit 150 MC (30W), receive 450 MC. Will operate in Ham Bands. No modification required. Large stock available. $150.00 each. Omni Communications. Call (312) 852-0738.

MOTOROLA VHF (2 meters) U43 Motran, 4 channel all solid state. Omni Communications. Call (312) 852-0738.

Coming Events

ACTIVITIES

"Places to go..."

ILLINOIS: Wheaton Community Radio Amateurs Hamfest will be held February 6, 1983 at Arlington Park Race Track EXPO Center, Arlington Heights, Illinois. Free Flea Market tables and plenty of floor space. Large commercial area including computer section. For general info call W9JTO at 311-231-9524. Clear paved parking. Awards. Tickets $3.00 at entrance, $2.50 in advance. Send SASE to WCRA, P.O. Box OSL, Wheaton, IL 60187. Talk-in on 146.01/61 and 146.94. Doors open 8 AM. Be There! — KA9KDC.

INDIANA — SOUTH BEND: Hamfest Swap & Shop, January 2, 1982, first Sunday after New Year's Day at Century Center downtown on U.S. 33 Oneway North between St. Joseph Bank Building and river. Industrial History Museum in same building. Carpeted half acre room. Tables $3 each. Four lane highways to door from all directions. Talk-in freq: 52-52, 99-39, 93-33, 78-18, 69-09, 145.43, 145.29.

LOUISIANA: The Southeastern LA University and the Southeastern LA ARC are sponsoring a Hamfest on Saturday, January 15 on the SLU campus, Twelve Oaks Cafeteria from 9 AM to 3 PM. Free admission. Free swap tables.

MICHIGAN: The Southfield High School Amateur Radio Club's 18th annual Swap & Shop, January 30, Southfield High School, 24675 Lahser, Southfield, 8 AM to 3 PM. Admission $2.50. Reserved tables $18.00 in advance for two 8 ft. tables. All profits go toward electronic scholarships and to support the activities of Southfield HS Amateur Radio Club. For information/reservations: Robert Younker, Southfield High School, 24675 Lahser, Southfield, MI 48034. Telephone: (313) 354-7372, 8 AM to 10:30 AM; or (313) 354-8210 10:30 AM to 3 PM, Monday through Friday.

VIRGINIA: Richmond Frostfest '83. The annual winter Ham Radio and Computer Show will be held Sunday, January 16 at the State Fairgrounds, Richmond. General admission: $4.00. All indoor flea market and commercial exhibits. Major prizes in HF and VHF equipment and a minicomputer. Sponsored by the Richmond Amateur Telecommunications Society, P.O. Box 1070, Richmond, VA 23208.

OPERATING EVENTS

"Things to do..."

FEBRUARY 5: New Hampshire QSO Party sponsored by the Concord Brasspounders, Inc. Operating periods 1900Z February 5 to 0700Z February 6 and 1400Z February 6 to 0200Z February 7. 24 hours total. Awards mailing deadline March 12, 1983. Send your entry with large SASE for results and/or award to Concord Brasspounders, Inc., c/o Norman W. Littlefield, RFD 1, Buck Street, Box 323, Suncook, NH 03275, W1JBX.

JANUARY 15: WD2ALL will operate the Camp Ballou Scout Freezeout from the Camp Ballou Boy Scout Camp; 1400 to 1700, 1800 to 2200, and 2300 to 0100 GMT. Frequencies 10 kHz above lower edge of General phone bands and 25 kHz above lower edge of Novice bands. Also 146.55 simplex operation is planned. QSL with SASE to WD2ALL via callbook.

HAMFESTERS RADIO CLUB is celebrating its 50th anniversary in 1983. Look for club station W9AA on all bands. We will send a special QSL card for contacts this anniversary year. QSL to: Hamfesters Radio Club, P.O. Box 42792, Evergreen Park, IL 60642.

JANUARY 29: The fifth annual *Freeze Your Arctic Off* sponsored by the Ford Tin Lizzy Club from 1700Z to 1700Z, January 29 and 30. Look for AD8R/8 on 7.275, 21.380 and 146.58. For a certificate QSL to Box 545, Sterling Heights, MI 48077.

JANUARY 29: The Eau Claire, WI, ARC will operate K9EC/9 during the National 70 meter Ski Jumping and Nordic Combined Championship from 1400Z to 2300Z, January 29 and 30. Frequencies: CW — 52 kHz up from bottom edge. Phone — 3980, 7277, 14282, 21382, and 28620. SASE for certificate to N9AIX, P.O. Box 201, Altoona, WI 54720.

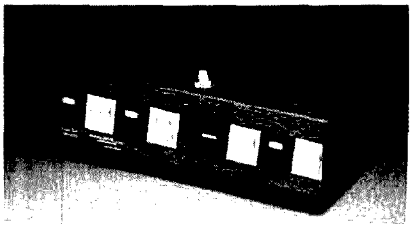
JANUARY 22: The West Virginia QSO Party sponsored by the WV State Amateur Radio Council, 1700Z January 22 to 1700Z January 23. Single operator only. Suggested frequencies: Phone 10 kHz from lower edge of General sub-bands; CW 35 kHz from low ends; Novice 35 kHz from low ends. Repeater contacts permissible. Mail logs by February 11 to K8BS, 950 Gordon Road, Charleston, WV 25303.

NEW products

surge protection and master control

Alpha Delta Communications has introduced the new Master AC Control Console (MACC) which features power surge protection and centralized control of several components. The MACC unit plugs into a single outlet, providing eight plug-in grounded outlets of its own — one hot, for a continuously powered application such as a clock and seven for individually controllable components.

The front panel has rocker switches for the individually controllable components, plus a master control rocker which allows the entire system to be turned on or off. Rockers are lit when on.



Alpha Delta's MACC uses three-stage automatically restorable circuitry to clip off power spikes and surges, and has a manually resettable circuit breaker for further protection. MACC is tested to IEEE pulse standards and rated at 15 amperes, 125 Vac, 60 Hz, 1875 watts continuous-duty total for the console.

MACC is priced at $79.95. For more information, contact Alpha Delta Communications, P.O. Box 571, Centerville, Ohio 45459.

X-Panda-Five

The X-Panda-Five makes your Hustler (or equivalent) into a five-band mobile antenna, with the appropriate coils added. When the X-Panda is installed with the proper resonators, you can change bands without stopping your vehicle. Each resonator, tuned to the desired frequency, eliminates the need for an antenna tuner.

Also, the X-Panda-Five adapter added to the Hustler or Hy-Gain antenna with the appropriate resonators and added ground planes will make an ideal antenna system for apartment houses and condominiums. It can also be used to make a multi-band antenna system for vans, campers, motor homes, and travel trailers.

The X-Panda-Five will accept either regular or super-size resonators. For further information, contact JL Industries, P.O. Box 030413, Ft. Lauderdale, Florida 33303.

new TS830, TS930, R820 filter kit

Fox Tango Corporation announces a special high-quality matched-filter kit designed to significantly improve the selectivity of the Kenwood R820, TS830, and the new TS930 series. The Fox Tango filters (both 8-pole discrete crystal units instead of the original monolithic and ceramic types) each have a bandwidth of 2.1 kHz (net bandwidth of 1.99 kHz); a combined shape factor of 1.19; and an ultimate rejection greater than 110 dB. VBT may be used to narrow the operating i-f bandwidth to reduce QRM, (narrower bandwidths, usually given at −6 dB, help reduce adjacent channel interference) but the steepness of the filter skirts or shape factor (−60 dB BW divided by −6 dB BW) and their depth (ultimate rejection) are more important. If VBT is used to reduce the bandwidth to 1.99 kHz (to equal that of the FT filters without VBT), the shape factor of the original filters becomes 1.45 as compared with 1.19 — and the ultimate rejection is less than 80 dB as compared with more than 110 dB — both significant differences.

Regardless of the type of filters, the use of VBT in these receivers always reduces the shape factor. With VBT off, the characteristic curves of the two filters essentially coincide with one another, referred to as filter cascading. The combined shape factor is usually better than that of either of the two filters involved. When VBT is used, one filter characteristic is made to slide with respect to the other, and only the portion where they overlap represents the bandpass area. The cascading effect is lost and the resulting characteristic has the skirt of the first filter on one side and that of the second on the other.

Also, because of the rounded shoulders of the original filter characteristics, the overlap at narrower bandwidths has the effect of increasing the filter insertion loss: 5 dB versus 0 dB with FT filters at 500 Hz bandwidth; 10 dB versus 1 dB with FT filters at 300 Hz. The greater such losses, the lower the receiver sensitivity in the CW mode. The superior characteristics of the FT filters results in excellent performance in both the SSB and CW modes practically eliminating the need for the purchase of optional CW filters by all but the most serious CW operators.

There are significant advantages in not buying any CW filters. In addition to saving the cost of the CW filters, installation is simplified since the FT matched pair can be inserted directly

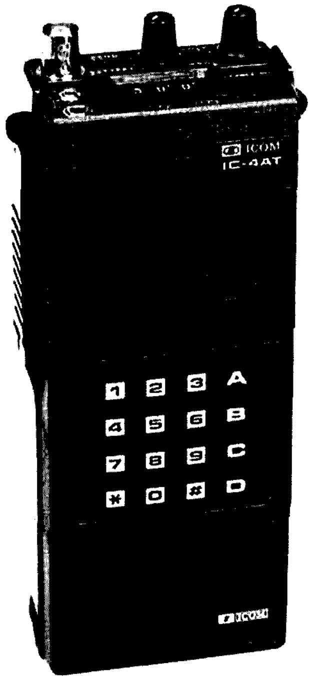
into the holes provided for the CW filters. With this arrangment, the following operating options become possible: (1) FT filters for both RX and TX; (2) FT for RX, original Kenwood for TX; (3) FT for RX, switch-select FT or Kenwood for TX; (4) Switch-select FT or Kenwood for RX/TX. If CW filters have been (or are to be) used, the recommended arrangement is to replace the original SSB filters with the FT 2.1 matched pair. In this case only option 1 is possible. This installation is easy, no drilling or switching is required, and all parts are provided in the kit.

The matched pair filter kit, complete with detailed instructions, two 2.1 kHz Fox-Tango filters (guaranteed for one year), and all needed cables and parts is offered at an introductory price of $150 plus $3 for shipping ($5 for air). Send your order, specifying the rig with which the filters are to be used, to Fox Tango Corporation, P.O. Box 15944, W. Palm Beach, Florida 33406, or order by telephone: 305-683-9587.

450-MHz handheld

A second cousin to the popular IC-2/3A series, the IC-4AT provides coverage of the 440-MHz band. It is identical in appearance, size and operational features to the popular IC-2A/3A series. All accessories, including battery packs, chargers, microphone, etc., are completely compatible with the IC-2AT and IC-3AT series. The IC-4AT also includes a sixteen-button Touchtone® pad.

The IC-4A covers the 440-MHz band from 440.0 MHz to 449.995 MHz and is set up for both duplex and simplex operation. The power output is nominally 1.5 watts with the standard IC-BP3. The IC-4A system will come complete with IC-BP3 Ni-Cd battery pack, wall charger, belt clip, rubber duckie, and wrist strap. The IC-4A costs $269 and the IC-4AT is $299.



For more information, contact ICOM America, Inc., 2112 116th Ave. N.E., Bellevue, Washington 98004; telephone 206-454-8155.

universal programmable filter

Applied Invention has introduced Reticon R5620, a universal programmable active filter. The R5620 is a complex NMOS switched capacitor active filter (SCF) analog IC. It uses switched capacitor technology to synthesize a two-pole pair active filter that requires no external components and operates over the range of 0.05 Hz to 25 kHz. The five basic filter types — lowpass, highpass, bandpass, band reject, and all pass — can be used as well as a programmable sine-wave oscillator. The Q is set to one of thirty-two approximately logarithmically spaced values from 0.57 to 150 by five control pins (hard-wired or TTL/MOS logic levels). Center frequency is set by an external clock oscillator. The clock division ratio can

be varied from 50 to 200 (two octaves) in logarithmic steps.

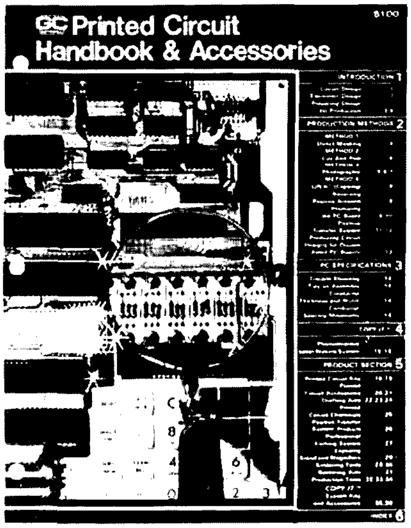
Switched capacitor filters are simply standard analog filters with fixed resistors replaced by time division variable-switched capacitors. This results in a highly stable filter which can be accurately tuned with only a variable clock source. The programmability of the SCF makes it very attractive for microcomputer-controlled synthesizers and other analog/digital system applications. Small quantity price of the Reticon R5620 is $8.00, with an application package available for an additional $2.00. Contact Applied Invention, RD2, Route 21, Hillsdale, New York 12520; telephone 518-325-3911.

new PC handbook

A new 40-page *Printed Circuit Handbook and Accessories Catalog* from GC Electronics, a division of Household International, features step-by-step instructions and diagrams and explains in careful detail how to produce professional-quality printed circuit designs.

This latest handbook is an enlarged and more complete version of the company's previous edition. It includes a variety of helpful information on how to produce both single and double sided printed circuit boards as well as PC specifications and trouble shooting tips. In addition, it pictures and describes more than 125 GC products.

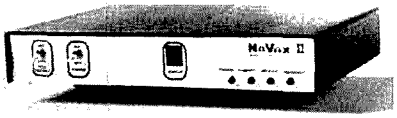
Copies of the new PC *Handbook* (Cat. No. 22-100-HB) are now avail-



able through GC Electronics' distributors nationwide. For more information, contact GC Electronics, 400 S. Wyman, Rockford, Illinois 61101; telephone 815-968-9661.

simplex/duplex autopatch model

Current Development Corporation announces the Novax II Mobile Connection for interfacing radios to DTMF (TouchTone) and rotary dial telephones. In addition to the standard features of Novax I, the Novax II offers a four-digit access code; LED display; toll restrict; repeater use; and interfaces to rotary dial systems (for rotary dial telephones). Both units use a high speed switching technique that eliminates voice-activated switching problems.



For more information contact Current Development Corporation, Box 162, Tudman Road, Westmoreland, New York 13490; telephone 315-829-2785.

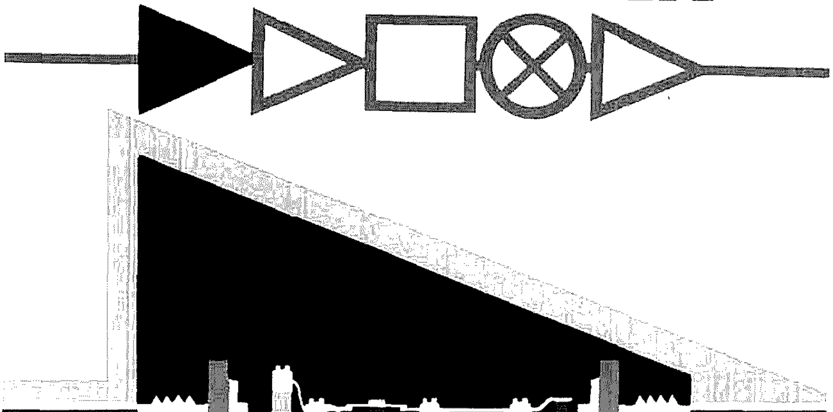
radio

- CB to 10 fm
- the Bragg-Cell receiver
- VFOs tuned by cylinder and disc
- the Bobtail curtain

hr

focus on communications technology

A LOW-NOISE 2304-MHz PREAMPLIFIER



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contents

REFLECTIONS

The Battlefield

It's 0600 hours UCT. Most sensible people on this side of the Atlantic have turned in for the night. Europe is waking up to a new day. The place is 75 meters and we are greeted with a mixture of howls that sound like the sound track from a horror movie. Upon closer examination human voices become recognizable, along with radioteleprinter signals, several unmodulated carriers, and the ever-present and pervasive noise.

During a brief lull in the hostilities, a crisp, British-accented voice is heard, announcing that he is listening for any stateside station, preferably Midwest or West Coast. His 10 over 9 signal attracts quite a bit of attention and the melee begins. As if the entire FCC roster were being called, one-by-twos, one-by-threes, and two-by-ones line up, each in their own turn, to shout their calls in the hope of attracting the British station.

A second pause, and a confirmation by the G station is heard. How delightful! His signal is strong, the band is wide open, and the noise level is down. But what's this? A 30 dB over S9 carrier sweeps back and forth, the work of a disgruntled Amateur who feels that he must get even — and, for most listening, he's accomplished his task. For them, he's turned what might best have been described as merely a headache-producing operating experience into one that sends the blood pressure up and poisons the bile. Others, however, accept it as inevitable, switch to their Beverage antenna, narrow their passband, and insert rejection. For those few, communications technology has moved on and they're riding with it.

Six years have passed since I left the East Coast. How simple it all seemed then. A trap vertical, four quarter-wave radials, and 25 kHz of band on which to meet our overseas partners leisurely and on an equal footing. The exceptionally well-equipped station had a pair of phased verticals, twenty radials under each, and maybe even a Beverage antenna for listening. Today, during *non*-contest periods, even that station will not necessarily produce a response on the first call. More and more we hear of the four-square (four phased verticals in a square configuration) with radial systems measured in miles not feet, 1200-foot rhombics on a leg, and three-element Yagis, most fixed, some rotary.

So what's the complaint? This is progress, isn't it? Perhaps it's the fact that this all occurs on 3799 ± 0.000 . . . 1 kHz; as if some magical gentleman's agreement has been made by the unseen multitudes.*

We must love this band. What other word could describe the rush of emotion while working DX, could explain the reason we put up with operating conditions that would make the military C^3 (Command, Control, and Communication) people wince, the sore muscles, the strained wallet, and lost sleep?

Is it possible — not in a world far, far away but right here, on the dial between 3777 (remember LSB) and 3800 — to improve our act, show a little more patience, cut back on the processing, and listen a little more? If not, I suppose I'll have to put in an order for 20,000 more feet of radial wire, 500 feet of six-inch irrigation tubing, solid-state commutating for my Beverage farm.

Might the hostilities, if not cease, perhaps slacken?

Rich Rosen, K2RR/1
technical editor

**SSB DXers from Europe, Africa, Asia, the U.S., and elsewhere have gravitated toward 3799 for several reasons. It's there that one finds the greatest commonality of nationally regulated Amateur frequencies and the least interference from worldwide commercial broadcast stations. Japanese Amateurs, for example, can operate only between 3793 and 3802, and Australians between 3794 and 3800, while Europeans and many others can go no higher than 3800 kHz. With strong commercial and military broadcasts from Regions 1 and 3 below 3795, and with three daily domestic nets in Region 2 also operating below 3795, 3797-3799 has become the 75-meter DXers' common ground.*

But what about the Extra Class Amateurs who don't want to chase DX and yet who operate in the "window"? They of course have every legal right to use it. If only they would bear in mind that there is a difference of up to 70 dB in signal level between their signals and those of the DX stations! That's quite a bit of filtering, directive antenna gain, and rejection that's needed to even come close to equalizing. Also, communicators are supposed to use the minimum necessary power to establish and maintain contact. Rarely is a kilowatt needed to communicate across town. By using a 10-dB-step power attenuator (from 0 dBm to +60 dBm in six steps), contact with several hams over 500 miles away has been accomplished at the 10-milliwatt level on 3795. If most Amateurs followed this rule (minimum necessary power) we might be pleasantly surprised by how much nicer 75 meters would sound.

comments

filters

Dear HR:

The FCC Rules and Regulations of January, 1979, Part 97, Amateur Service, state that the second harmonic from an Amateur transmitter must be down from the fundamental by 40 dB. The responsibility to comply falls directly upon the Radio Amateur. But few have the equipment to make this measurement accurately, and even the inspectors who recently cited a local Amateur had to first obtain the equipment to make the measurement.

One might assume that an AB-1 or AB-2 amplifier is operating linearly if it is not being over-driven and the bias is set right. One might assume that the lowpass filter will take care of the problem — it's easy to forget that it cuts off at 30 MHz, and that your station might be measured at 7 MHz.

The best solution for the Amateur is to use a band cut-off filter as described by *GE Ham News*, June, 1957, Vol. 12 No. 2. The mica capacitors, however, are located in the form of an inductance, which throws the cut-off frequency off. If you use the small ceramic type, any high power will make popcorn out of them. The only capacitors I have found to be satisfactory are the Ceramic TVL Centralab, but they are expensive. They cost $10.00, and you need four per band.

Amateurs do not realize that the FCC is *not* talking about TVI; they are talking about the second harmonic output from the 160-80-40-20 meter bands. No one seems to care and the regulation is being ignored — except when an Amateur is cited and finds that he has to get a letter of certification from someone saying that his rig is OK. That's a service and shipping cost of several hundred dollars. All I can say to those who do not have a filter is, good luck!

Ed Marriner, W6XM
La Jolla, California

autodialer

Dear HR:

A source for drilled and plated printed-circuit boards for my August, 1982, *ham radio* article, "A Portable Touch-Tone Autodialer," is now available. Dynaclad Industries, P.O. Box 296, Meadow Lands, Pennsylvania 15347, will make them available at $8.00 per board plus $1.50 shipping.

Alan Lefkow, K2MWU
Thiells, New York

ultimate tone information

Dear HR:

In regard to my article, "The Ultimate Tone Decoder," in the September, 1982, issue of *ham radio*, please note that our chips were purchased from Seiger Associates, 1885 Hicks Road, Rolling Meadows, Illinois 60008. Also note that the capacitors connected to pin 2 and pin 11 of the 8865 chip are 20 picofarad capacitors.

E.M. Dean, WD9EIA
Machesney Park, Illinois

Q signals

Dear HR:

I don't know where the idea that the Q signals are only for CW got started, but I've seen it many times, as in the N4AGS letter in the April issue. The facts follow:

Q signals are a part of the International Radio Regulations, a multi-partate treaty signed by the U.S.A. They are set forth in Appendix 13 (1968 edition). Section I, paragraph 1, specifies that the signals QRA to QUZ are for the use of all services. (QAA to QNZ are for the aeronautical service and QOA to QQZ are for the maritime services.)

A useful exercise is to look up the meaning of QRJ, QSU, and QUE. Note also the phrase, in Appendix 13A, Section I, paragraph 3: "in radiotelephony spoken as CHARLIE or NO." And further, in Section II: "When used in radio-telegraphy a bar over the letters composing a signal denotes that the letters are to be sent as one signal," as in $\overline{AS}$, wait.

With respect to the use of a phonetic alphabet, the Radio Regulations, Appendix 16, paragraph 1, specifies that the Alfa, Bravo . . . phonetics shall be used when necessary to spell out call signs, service abbreviations and words. Amateur regulations are at variance with this, however, in that paragraph 97.84(g) only "encourages the use of a nationally or internationally recognized standard phonetic alphabet."

There are important practical reasons behind the International Regulations. The Q code is the same in all languages. Consequently, a real QSO can be completed without the participants knowing a word of each other's language. It's easy on phone, too — the International alphabet words were selected to be easy to pronounce and hear in most languages.

I have called on the ARRL to work to correct the current misuse which is so common (see the correspondence column, August, 1981, *QST*, page 61).

R.P. Haviland, W4MB
Daytona Baach, Florida

presstop de W9JUV

THE WARC 79 TREATY WAS FINALLY RATIFIED by the U.S. Senate Tuesday, December 21. Though ratification won't have any immediate effect on the U.S. Amateur community, it does mean that the FCC can now begin the regular rule-making procedures leading to permanent assignment to the Amateur service of the new 10, 18, 24, and 902 MHz WARC bands.

A "NO-CODE" AMATEUR LICENSE is almost certain to be proposed by the FCC early this year, very likely by the time this Presstop sees print. The alternatives still seem to be either dropping the code requirement from the current Technician license or adopting a new "digital" class license, such as is offered in Canada.
Within The FCC The Modified Technician License probably has the strongest support, as it would cost the least and require little staff effort to implement. It's also the course most vehemently opposed in the Amateur community, since it would not only permit individuals with no CW capability at all to become Amateurs but would also give them access to the HF bands. It is possible that some form of CW capability could still be required before a "no-code" tech could legally operate on an HF band, for example a "certification" by a General Class or higher Amateur that the individual can send and receive Morse code.
But The Strong Opposition To A No-Code Tech License already demonstrated by the ARRL and many individual Amateurs may very well lead the Commission to lean toward the "digital" type license, with a difficult technical exam like the Canadians'. Since that license has not proven popular in Canada, it could very well flop here too.
The Present FCC CW Tests May No Longer be nearly as effective a part of the Amateur exam process as most Amateurs believe. Some FCC Field Offices report an increasing number of applicants have apparently memorized the answers to at least one of the CW exams. They sit through the transmission, then answer the questions—though sometimes the answers will be for a tape other than the one they just heard! When lucky they pass the CW exam; if not, they return every month until they do.

THE 900-MHZ "PRIVATE RADIO COMMUNICATIONS SERVICE" is also quite certain to be proposed in a January Notice of Proposed Rule Making. Latest stories out of Washington say it won't be just a UHF CB service, but more a land mobile service readily available to anyone wishing to use it. Amateur Radio will not be connected in any way with the new service, despite some earlier rumors that it might.
Amateur Access To The New 902-928 MHz WARC Band seems to have received very little attention as yet, though signing of the WARC Treaty may now give it a push forward.

CORDLESS TELEPHONES ARE MAKING QRM for some 80-meter CW operators. The telephones use 1750 kHz for one side of the two-way circuit, and harmonics of the 1750-kHz carrier-current transmitter fall right into the low end of 80 CW. Though not confirmed, it's also likely that some 160-meter Amateurs are causing problems to neighbors' phones.
The Other Side Of The Phone Circuit, 49 MHz, may soon also be creating similar difficulties. Some makers of cordless phones have petitioned the FCC to extend their band edge up to 50 MHz, and with the questionable quality of some consumer electronics it is likely that phone interference to and from 6-meter users may become a problem as well.
Some "Long-Range" Cordless Phones, brought into the country by travelers or even smuggled in by people with little interest in or knowledge of frequency use, have also been showing up. These units operate at VHF and run considerable power. One, whose frequency capability includes the 2-meter band, promises 100-km range with an appropriate antenna and accessory power amplifier! Anyone hearing such a unit in operation should alert the nearest FCC monitoring station immediately.

RUSSIA'S ISKRA 3 SATELLITE HAS APPARENTLY FAILED for good. Though its beacon and 15 to 10 meter transponder were both heard early in December, by the middle of the month the over-temperature problem that had plagued the new bird seems to have shut it down.
Better News Is The Apparent Coming To Life of RS1, one of the two Russian satellites launched back in 1979. Several listeners have reported hearing noise and signals in its 29.35-29.40 MHz downlink passband, and at least a couple of satellite-relayed contacts have been made on RS1's frequencies when none of the other active birds were accessible. Its 29.4 MHz beacon has not been heard and is probably not functioning.
Asia Thanks To UA0BBN Has Been Worked by WØCY and W8DX via RS6. Though the window to the Siberian station is short, it does provide a rare satellite WAC opportunity for many.

"AMATEUR RADIO'S NEW FRONTIER" is the tentative title of a new Amateur Radio video tape promotion that's to be produced by a group headed by Roy Neal, K6DUE. Sponsored by the ARRL, the new effort will be directed at teenagers and will emphasize space-age communications and computer technology. Locales for the production will include Johnson Space Center in Houston, Kennedy Spaceflight Center, and AMSAT and ARRL headquarters. After funding is approved by the League directors, it will be targeted for completion by September.

THE FEDERAL JUDGE HEARING THE BURBANK TOWER CASE has agreed to take under advisement the city's motion to dismiss the suit filed by Burbank Amateurs. How soon he'll rule on the motion against that suit cannot be predicted at this time. The suit seeks to overturn the city's ordinance prohibiting new tower construction and outlawing RFI.

a low-noise preamplifier for 2304 MHz

0.8-dB noise figure and 16-dB gain in a home-built microwave amplifier

Even though commercial equipment is not available for the 2300-2450 MHz band, weak-signal operation at 2304 MHz is undergoing great growth: The W2SZ/1 contest station worked only four other stations in four sections (including one station in eastern Pennsylvania, over 250 miles away) on this band in the June, 1981, VHF contest; in the June, 1982, VHF contest, eleven stations in eight sections were worked (including two stations in eastern Pennsylvania). Additional stations were known to be available, but were not worked because of a transmitter high-voltage relay problem.

The first requirement for a 2304-MHz station is a receiving converter. Many designs exist.[1] If low-noise preamplification and proper filtering are used in front of a subsequent mixer, then even a simple 3-dB hybrid-coupler mixer,[2] etched on a G-10 printed circuit board and using low-cost Schottky diodes (such as the HP 5082-2810), is adequate. **Fig. 1** shows a block diagram of a receiving converter, along with stage noise figures and gain/loss values. The 27 dB of gain in front of the mixer is more than adequate to overcome the 8-10 dB noise figure of the mixer. The bandpass filter[3] is used to reject noise and image signals at the mixer image frequency, aiding the relatively broadband preamplifiers. The i-f amplifier[4] may or may not be necessary, depending on the sensitivity and noise figure of the receiver at the selected i-f.

Because all these requirements have been previously discussed in other articles,[5] as have been the formulas necessary to obtain the overall converter gain (about 40 dB) and noise figure (approximately 1.0 dB), this article concentrates on the low-noise GaAs fet first-stage amplifier. The rf amplifier used in the second stage is a well-known microstrip design[6] using a bipolar NE64535 transistor (which costs about $7).

The device selected for the LNA is a Mitsubishi MGF-1402 GaAs fet, presently priced at about $15. The LNA circuit is shown in **fig. 2**. On the basis of past experience, I selected a π section to impedance-match the device output to a 50-ohm load. The length and width of output inductor L2 are determined by the required inductance and the height of the supporting portion of the π network tuning capacitors C3 and C4. The best information available, at the time, indicated that the optimum noise impedance to be presented to the gate of the device is between about 85+j60 ohms and about 110+j90 ohms. The input circuit was designed to provide an acceptable range of impedances, around these desired values, to accommodate variations between devices. Source self-bias is used; effective series-resonant chip bypass capacitors (C_S) are an absolute necessity. Fortunately, a set of five chip capacitors are available, at a reasonable price, from the same source[7] that supplies the GaAs fet device. These chip capacitors, the variable capacitors, and the device itself are all relatively small. Use of sharp-pointed tweezers is advisable for careful handling of these parts. While only four chip capacitors are needed, the fifth chip capacitor is insurance, as the little beasties are easily destroyed or lost. I did all soldering

By Geoff Krauss, WA2GFP, 16 Riviera Drive, Latham, New York 12110

with a 23-watt pencil line, while observing special grounding techniques for handling the device.

construction

The LNA is built from the output forward. Refer to **fig. 3**. A base of copper-clad printed circuit board is cut to a width of about 1-1/2 inches (40 mm) and a length in excess of 3 inches (80 mm). End plate 1 is formed from a piece of double-sided printed circuit board, cut to a width of 1-1/2 inches and a height of 1 inch. Only the outside copper cladding of the end piece 1 is initially soldered to the base piece. A hole is now formed in the end piece 1 to pass the threaded portion of a gold-flashed, square-flange SMA connector, J2. The hole is positioned so that the edge of the connector's flange rests on the surface of the base, inside the angle formed by the base and end piece 1, when the threaded portion extends through the end piece hole. The gold flashing on the connector readily accepts solder, allowing the flange to be soldered to the end plate and base plate along all four edges with a minimum of heat. The pin of the connector lies along the center line of the base — almost all of the components are mounted along the center line. The rest of the inner angle between the end piece and the base is soldered after J2 is installed.

The output tuning capacitors C3 and C4 are mounted next. First, form the C4 lead nearest to the tuning screw to extend over the output connector J2 center pin. A chip capacitor, C5, is placed between the pin and the lead at a later time. When the lead is formed and properly placed, solder the base of tuning capacitor C4 to the copper foil of the base. Use of a silver-bearing, low-temperature solder and appropriate flux is highly recommended for soldering *all* components, and is a must for chip capacitor soldering. The small circular formations about the screw end of capacitors C3 and C4 provide buttresses upon which output inductance L2 is later mounted. Therefore, solder capacitor C3 in place along the center line, at a distance from capacitor C4 such that it can receive the strip inductor, and with the C3 lead pointing along the center line and away from the inductor position.

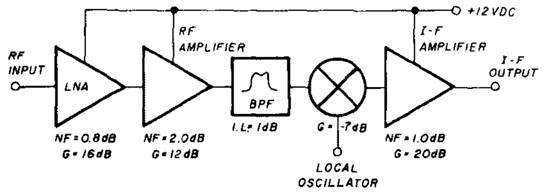

fig. 1. Block diagram of a receiving converter.

After C3 and C4 are mounted, solder inductor L2 between the variable capacitors. This soldering to the capacitors should be carefully done, and preferably from the underside of the inductor, to prevent solder from flowing into the tuning screw mechanism of the capacitors. *Be aware that, if different capacitors are used for C3 and C4 with different buttress heights above the copper groundplane of the base, the width of inductor L2 may have to be adjusted to compensate.*

Next, form the X support from a piece of copper plate or foil, as shown in **fig. 5**. Mount the two source lead chip capacitors (C_S) on the top of the support, with their inner surfaces about 1/8 inch (3 mm) apart. The height of the X support is such that the top of the chip capacitors is approximately in a plane with the leads of capacitors C2 and C3, when the support is soldered to the copper base covering. When the support has been properly formed and placed, solder the lower tabs A to the base.

Now move the input shunt capacitor, C2, into place, with its lead extending along the center line toward the output connector; trim the lead to extend along a line between the two chip capacitors on support X. The distance, D, between the center line of

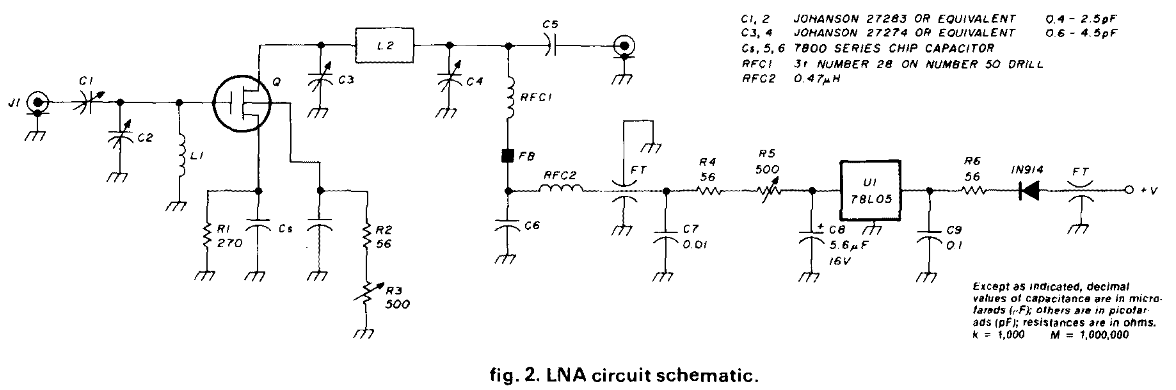

fig. 2. LNA circuit schematic.

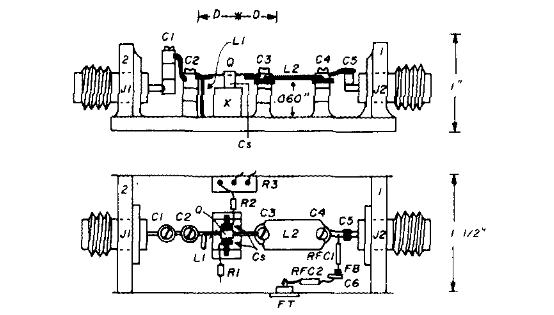

fig. 3. Construction details of the low-noise amplifier.

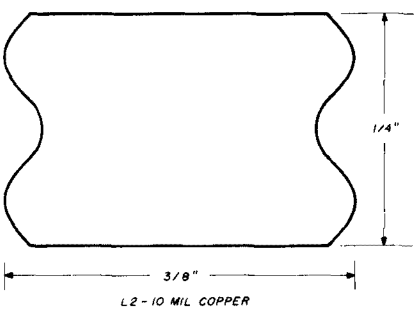

fig. 4. Output pi-inductance, L2.

the chip capacitors and the center lines of each of capacitors C2 and C3 should be between about 3/16 of an inch (4.5 mm) and 1/4 inch (6.5 mm). Solder the base of capacitor C2 in place. Now solder the base of capacitor C1 to the center pin of the input SMA connector J1, and form the lead from the other end of C1 over the shortest possible distance to the top of capacitor C2. Place connector J1 and capacitor C1 along the center line and tack the connector flange to the base.

At this point, make up another end plate with a hole to pass the threaded end of J1. Place end plate 2 over J1, solder the outside foil of end plate 2 to the base, and then solder the four flanges of connector J1 to the inner foil of the end plate and the base. Now solder the top lead from capacitor C1 to the top of capacitor C2. Form the input inductor, L1, as shown in **fig. 6** and solder from the top of capacitor C2 to the base foil.

The output chip capacitor, C5, is now soldered between the C4 lead and the output connector center pin. This is most easily accomplished by pre-tinning the center pin and soldering one end of chip capacitor C5 to that pin, before pressing the C4 capacitor lead down onto the other end of chip capacitor C5 for soldering. Add choke RFC 1, the ferrite bead, and the chip bypass capacitance C6 (having one end soldered to the base plate).

A printed-circuit-board piece is now added to each side of the base after forming a hole in one side for the feed-through capacitance, FT. After the sides have been added, solder in capacitor FT and RF choke RFC 2 between the feed-through and chip C6 capacitors. Solder the 270-ohm, 1/8-watt resistor R1 between ground and the free end of one of the source chip capacitors, C_S. Mount variable resistor R3 and then connect the 56-ohm R2 to the free end of the remaining source chip capacitor. The drain supply network of C7-C10, R4-R6, U1, and the 1N914 diode, can be mounted outside the LNA box (either on the surface or in a separate box section), but with no connection between U1 and R5. Only device Q remains to be mounted. While carefully holding one of the source leads with a grounded tweezer, use a low-wattage, grounded soldering iron to solder each of the source leads to the associated chip capacitor. The full length of the source lead is allowed to remain, as it serves as a convenient connection point for measuring bias voltages. Carefully cut the drain and gate leads to size and solder to the leads of capacitors C3 and C2, respectively. Construction is now complete.

tune-up

Adjust resistors R3 and R5 for maximum resistance. Apply a voltage, between 8 and 15 volts, to the power input and check for 5 volts at the output of integrated circuit regulator U1. After checking, connect the regulator output to variable resistor R5. Connect a voltmeter from ground to one of the source leads and, after again applying power, note a positive voltage of between 0.5 and 1.3 volts. Apply a relatively weak (less than −30 dBm) signal to input connector J1 and monitor the output signal at connector J2. Adjust resistor R3 for maximum gain, while adjusting resistor R5 to keep the drain source voltage (measured between a source lead and the top of chip capacitor C6) between 2.5 to 3.0 volts.

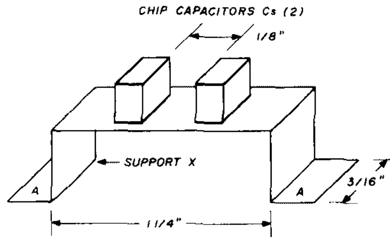

fig. 5. The X support, constructed from a piece of copper plate or foil.

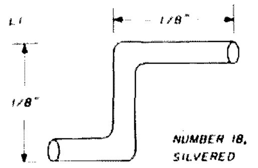

fig. 6. Input inductance, L1.

Now adjust capacitors C1 through C4 for maximum gain. Adjustment of drain current (with resistor R3) and drain voltage (with resistor R5) can be touched up for maximum gain. Using this maximum gain tuning procedure, a gain of about 20 dB with a noise figure of less than 2 dB is obtained. For minimum noise figure (measured to be about 0.8 dB, with an associated gain of about 16 dB), a noise source or weak-signal tuning method must be used. Do not change the tuning of output capacitors C3 or C4, but tune only input capacitors C1 and C2 for minimum noise figure or best weak-signal-to-noise ratio.

conclusion

A low-noise, high-gain amplifier for the 2304-MHz band can be built with a noise figure of under 1 dB for a cost less than $50 (dependent upon the state of your junk box). Outstanding reception is therefore possible on the 13 cm band.

See you there, next contest?

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ham radio

CB to 10 fm
one group's approach

An ingenious conversion that will quickly put you on the air on 10 meters

Recent articles in the ham journals,[1,2,3] regarding the conversion of CB rigs and circuit boards to 10-meter fm have caused quite a stir of activity. Here in the Toronto area, 10 fm is growing daily.

Our group decided to get on 10 fm by going the surplus route. Circuit boards made by the Japanese Cybernet company for Hy-Gain and several other manufacturers seemed to offer the best promise. They are essentially complete transceivers except for the addition of volume and squelch controls, microphone, speaker, channel selector switch, and housing. Several boards and forty channel switches were purchased from a surplus outlet,[4] crystals were ordered, and construction commenced. VE3FIT was fortunate enough to pick up a used forty-channel Hy-Gain CB rig for $10. This unit became the test bed.

initial set-up

First, make a visit to your public library and take a look at a copy of *Sams Photofact #148*. This book covers several of the Hy-Gain CB rigs that use the Cybernet PC board. Wire the controls and connections mentioned above as per the Sam's schematics. At this point it is probably best to leave off features such as RIT and ANL and concentrate on getting the basic rig operational on the CB channels.

Locate IC 101, the phase lock loop (PLL) chip. The leads from the channel connect to IC 101, but for now we'll just hard-wire them. Make sure that pins 8, 9, and 10 are joined, floating free of any components. Likewise for pin 7. Pick up +5 volts from pin 1 of IC 101 and temporarily jumper it to pins 7 and 14 of IC 101. Don't worry about the other input pins of the PLL chip. They have on-board pull-down resistors. This will program the rig to an output of 27.305 MHz (CB channel 30).

Go through the transmitter section, peaking each coil in turn. A General Cement Electronics alignment

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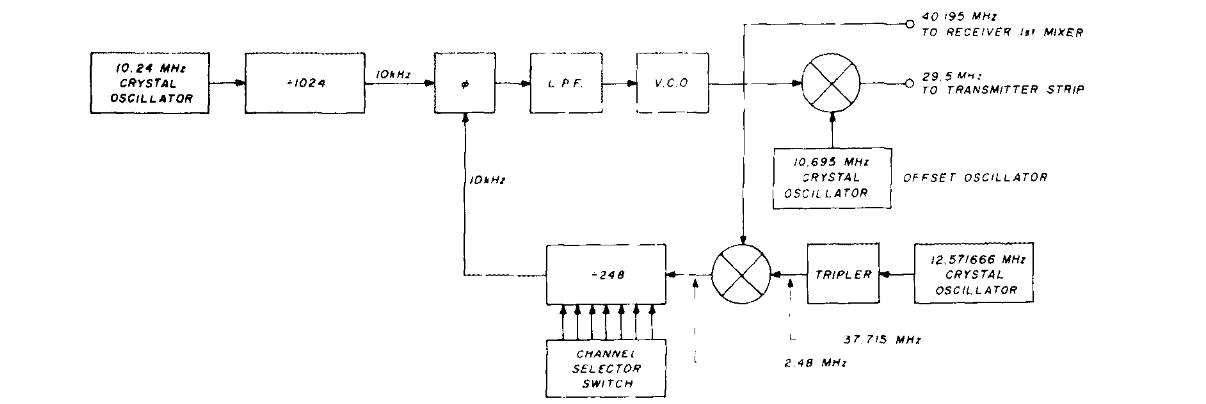


fig. 1. Mixing scheme.

tool kit (#18-530) is all you need. Use a #47 pilot lamp as a dummy load. If things are out of alignment, a general coverage receiver with an antenna close to the CB rig will provide an excellent output indicator. Simply tweak for maximum S-meter indication on the receiver.

The Cybernet receiver section is essentially prealigned. You may, however, wish to peak T104 and T105 in the rf amplifier stage. Use a signal generator as a signal source, or peak on a local CB conversation. Besides setting up the receiver, this will remind you of why you're glad to be a ham.

Once you are satisfied that the transceiver is operational, it's time to begin the conversion to Amateur use.

transmitter modifications

Remove crystal X101 (11.806 MHz) and replace it with a unit specified at 12.571666 MHz, HC18/U holder and 30 pF load capacitance. Notice the 6s at the end of the frequency. Leaving them off could ultimately put you as much as 2 kHz off frequency.

With the crystal now changed, power up the board and listen for the transmitter output on your main station receiver at 29.6 MHz, the center frequency of fm activity. Adjust T101 until the voltage at TP8 reads +2.1 volts. Peak T111, L104, T102, T103, L106, L109, and L110 for maximum output as shown on the lamp dummy load.

Once it had been "tweaked and peaked," our rig put out enough power to light a #47 lamp dummy load to about half brilliance. But prior to conversion, while still on CB, it had been very bright indeed! Obviously something was amiss.

The rig had originally been equipped with a pi-section lowpass filter composed of C604, C605, (330 pF each), and L600 (0.18 μH). Unfortunately, this combination cuts off somewhat below the fm operating frequencies. When the pi net was removed and replaced with a piece of wire and a good outboard lowpass filter, output increased noticeably but was still far short of the original level. The original three-component design just doesn't have enough harmonic attenuation to justify reworking it for 10 meters.

Next, C603 was reduced from 220 pF to 200 pF and L109 and L110 were readjusted for maximum output. This helped a bit.

Luckily, enough voltage measurements had been taken at various points in the transmitter stage (while still on CB) to enable us to determine where we were losing out. These readings are given in **table 1**.

table 1. Scope readings.

test point			waveform (*)
Q101	base		0.10
	collector		0.35
Q105	emitter		1.20
Q108	base		0.10
	collector		0.20
Q109	emitter		1.00
Q110	collector		1.00
Q111	base		0.20
	collector		4.00
Q112	base		1.90
	collector		8.50
Q113	base		2.00
	collector		17.0
ant.	connector	across 50 ohms (loaded down by probe)	26.0

*All voltages are ac peak-to-peak and measured with a high-impedance scope probe unless otherwise noted.

table 2. Crystal oscillator test points.

oscillator frequency	test point
10.24 MHz	pin 3, IC 101
12.571666 MHz (X3 = 37.715 MHz)	emitter Q105
10.695 MHz	emitter Q109

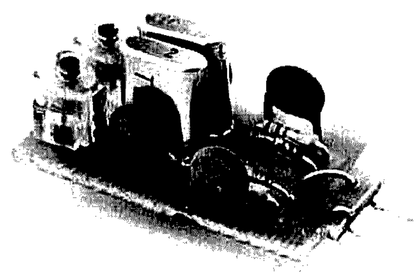


fig. 2. PC board making use of diode switching.

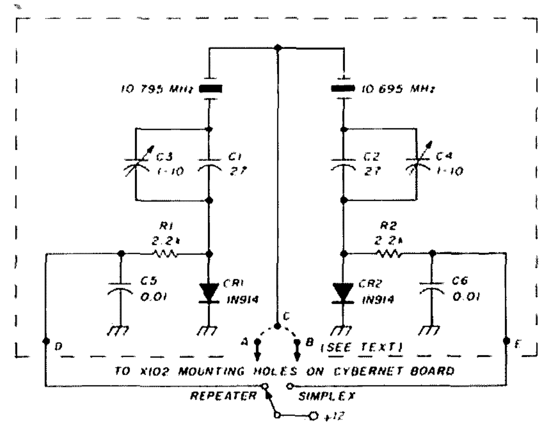


fig. 3. Repeater/simplex selector.

For maximum output, L106's slug had to be completely removed from its windings: that is, it was at minimum inductance and wouldn't go any lower. Removing a turn or two from L106 was impracticable because of the way the coil is built. To solve this problem, we simply broke one of the circuit paths from L106 and added a 270-pF capacitor in series. This change permits L106 to properly match Q112 and Q113. Setting L106 was now possible. We were getting closer.

Ultimately, the trouble was traced to the base of Q110, where the VCO and offset oscillator signals are mixed. The offset oscillator output level hadn't changed, and so it was assumed that, for whatever reason, the VCO level had changed. Changing C136 from 150 pF to 82 pF restored full drive to Q110 and to the rest of the transmitter chain. A full 5 watts of power was now available at the transmitter output. By the way, the output power is very sensitive to supply voltage. We normally run the rigs with 13.5 volts.

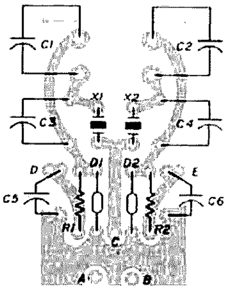
crystals

On our boards the crystals were all presented with too high a load capacitance. That is, C118, C127, and C178 were all 56-pF ceramic disks. A check of the oscillator output frequencies (at the points shown in **table 2**) confirmed that all crystals were oscillating slightly low in frequency. Changing the capacitors to 33 pF brought all the oscillators very close to the correct frequencies. If you can install trimmer capacitors, so much the better.

The mixing scheme used on the Cybernet boards is shown in block diagram form in **fig. 1**. Note that the rig is shown receiving and transmitting on 29.6 MHz simplex. The receiver's local oscillator signal is 10.695 MHz above the transmit frequency. In the repeater mode (input 100 kHz below repeater output), the receiver LO frequency stays the same but the transmitter frequency is now mixed down another 100 kHz.

Our group is contemplating repeater operation, so a PC board was designed to provide diode switching of X102 between 10.695 MHz (simplex) and 10.795 MHz (repeater). This board is shown in **figs. 2, 3,** and **4**. Diode switching was used because of the long distance between X102 and the nearest point on the front panel (about 5 inches). In the testbed rig, the ANL switch was used to provide the simplex/repeater selection.

This circuit board is installed vertically in the holes



(Ed note: D1 = CR1, D2 = CR2)

fig. 4. Offset oscillator parts layout.

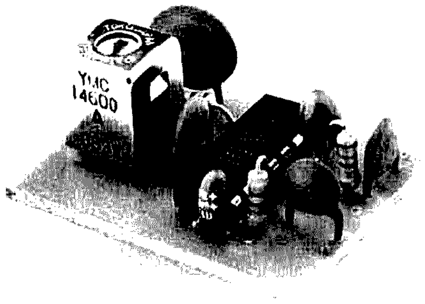


fig. 5. The fm demodulator board.

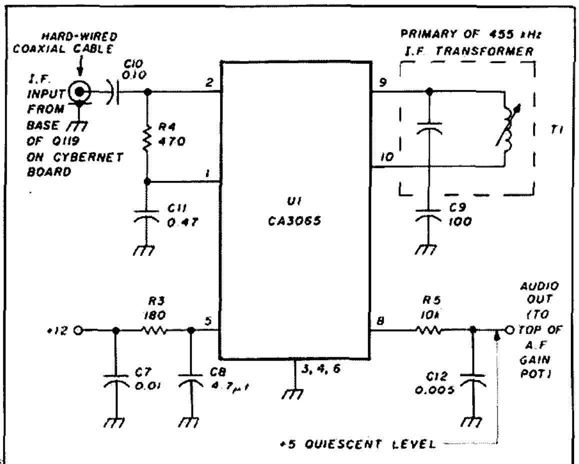


fig. 6. Fm demodulator.

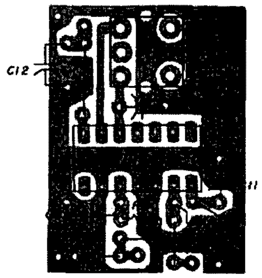


fig. 7. Fm demodulator board.

provided for X102 and secured with a drop of epoxy. Remove C127 and jumper its PC board holes. The lead coming from the common point of the two crystals, point C, can be jumpered to either terminal A or B. If A is the hot lead, ground B. If B is hot, ground A. This allows you to mount the board with the components facing inward, where you can adjust them. One of the subtle differences between the Hy-Gain and Cybernet boards is the placing of X102 and C127.

fm demodulator

The fm demodulator board shown in **figs. 5, 6,** and **7** is extremely compact and has been designed for operation at 455 kHz. A solder lug is soldered to the ground plane. This solder lug is then secured to the Cybernet board with a 4-40 (M3) screw and nut through the uppermost mounting hole in the audio amplifier's IC heatsink. Be sure to use shielded cable or Subminax (RG174/U) to feed the demod board from the base of Q119. Believe us, it's necessary! The bottom of the volume control is grounded and the wiper goes to pin 21 on the Cybernet board.

The i-f transformer appears to be a common Japanese transistor radio item. Ours measures about 0.4 × 0.4 inch (10 × 10 mm) and has a yellow core. It cost 50 cents at a local surplus store. The primary inductance is variable between 500 and 900 μH, and it resonates with an extremely tiny 180-pF tubular capacitor contained within the base of the i-f can. Initially, set the core about two turns from the maximum counter-clockwise rotation. Tune in a fairly weak signal (off the air or from a signal generator) and adjust the core for maximum undistorted audio output. This point is also coincident with maximum a-m rejection.

The CA3065 chip[5] contains an extremely sensitive i-f amplifier-limiter (200 μV for limiting), a differential peak detector (demodulator), and an audio output buffer. It will deliver over 4 volts peak to peak of clean audio. The demod section is shown in **fig. 8.** This circuit has also been referred to as a time delay differentiator.[6] What happens is that, at resonance, the tuned circuit impedance becomes purely resistive. This is shown in **fig. 8** as R. For our i-f transformer, R turned out to be about 70 kilohms. R and Cp must provide a phase shift of approximately −90 degrees. The required value of Cp at 455 kHz is about 100 pF. The output of the demod board goes to the top of the volume control.

fm modulator

To convert the rig to fm is fairly simple. The a-m

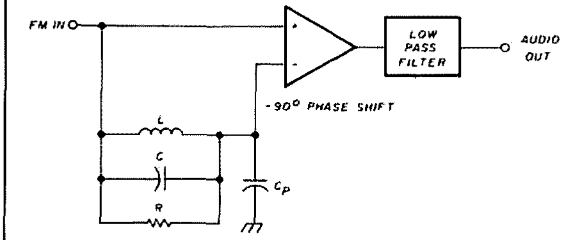


fig. 8. The fm demodulator, principle of operation.

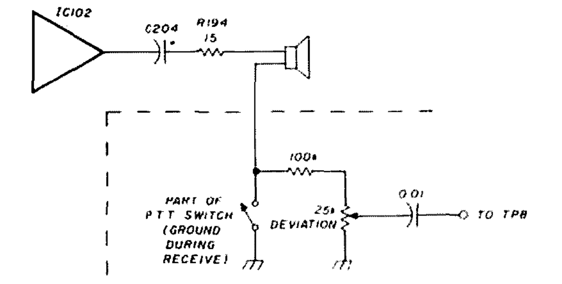


fig. 9. Frequency modulator.

modulation transformer T110 and RV102 are discarded. The B+ is restored to the final amplifier by joining points 18 and 20 on the Cybernet board. The output of the audio amplifier, IC 102, is routed from the positive side of C204 through R194 to the hot side of the speaker. The cold side of the speaker is grounded during receive through one of the contacts of the PTT switch.

During transmit, the amplified microphone audio is available at the cold side (now floating) of the speaker. The circuit shown in fig. 9 taps off part of this audio and applies it to the VCO control voltage at TP8, thus frequency modulating the signal. Setting the rig's deviation is done by monitoring the transmitter output on the general coverage receiver (a-m mode) using slope detection. Adjust the deviation control until clean audio is heard. On-the-air comments have been positive. If a deviation meter is available for this adjustment, so much the better.

forty-channel switch

If you were fortunate enough to obtain the forty-channel switch made for the Cybernet board you can use it directly. This switch is 1-1/2 × 1-1/2 × 5/8 inches thick (37 × 37 × 15 mm) and has a small printed circuit board on top. It's manufactured by Standard Grigsby. You may wish to reprogram the switch to another band plan, as will be described later. There is a very similar switch, also made by Standard Grigsby but without the PC board on top. This switch is meant for use with a circuit completely different from ours, and is almost useless to us. It also happens to be the switch we had purchased with our boards at three dollars a shot and were absolutely determined to use. Fortunately it can be modified to our specifications.

The switch is a marvel of mechanical design. It consists of two sections, a front-mounted detent mechanism and a rear-mounted printed circuit switch (see fig. 10). The sections are held together by two metal retainer pieces which are inserted from the rear. To remove them, slightly crimp the two metal tabs on the front of each retainer. Then, using a pair of pliers, pull the retainers out from the rear. The two sections will now separate.

The printed circuit section is held together by a bead of fairly soft epoxy, which also secures the terminals. Chip away at the epoxy (watch those terminals!) until you can separate the two halves. This will expose the removable printed circuit switch disk. This disk provides the coding to the PLL via ten finger contacts riding on the PC traces. The board etching dictates which fingers make contact with the common line (giving a digital "1"). This disk will be replaced with one custom-encoded to our needs, as discussed in the next section. Reassemble the sections and check for smoothness of operation.

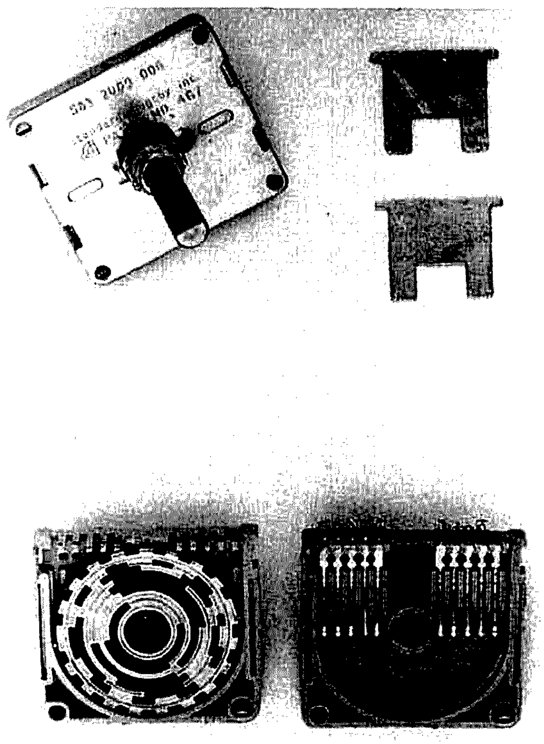


fig. 10. Forty-channel switch.

table 3. Wiring directions.

switch pin #	IC 101 pin #	division ratio
1	15	1
2	14	2
9	13	4
3	12	8
8	11	16
4	8,9,10	224
7	7	256
5,6	1	common (+5)

For the sake of convention we have numbered the switch terminals from 1 to 9, left to right, when viewing the switch assembly from the front. Wire these terminals to the appropriate points on the Cybernet board as per table 3.

The switch provides the PLL chip with a seven-bit word representing the division ratio necessary to place the transceiver on the desired frequency. The switch is directly connected to the PLL inputs to control the division factor. Table 4 gives this information. In the example shown, illustrating the mixing scheme (fig. 1), the channel selector switch provides the correct code to the PLL for division by 248.

Fig. 11 illustrates the internal finger contact arrangement of the switch. Note that each adjacent connection makes contact with every other switch

table 4. Division ratios.

channel number	switch pin # 7483921	division ratio	frequency MHz
01	0100100	228	29.300
02	0100101	229	29.310
03	0100110	230	29.320
04	0100111	231	29.330
05	0101000	232	29.340
06	0101001	233	29.350
07	0101010	234	29.360
08	0101011	235	29.370
09	0101100	236	29.380
10	0101101	237	29.390
11	0101110	238	29.400
12	0101111	239	29.410
13	0110000	240	29.420
14	0110001	241	29.430
15	0110010	242	29.440
16	0110011	243	29.450
17	0110100	244	29.460
18	0110101	245	29.470
19	0110110	246	29.480
20	0110111	247	29.490
21	0111000	248	29.500
22	0111001	249	29.510
23	0111010	250	29.520
24	0111011	251	29.530
25	0111100	252	29.540
26	0111101	253	29.550
27	0111110	254	29.560
28	0111111	255	29.570
29	1000000	256	29.580
30	1000001	257	29.590
31	1000010	258	29.600
32	1000011	259	29.610
33	1000100	260	29.620
34	1000101	261	29.630
35	1000110	262	29.640
36	1000111	263	29.650
37	1001000	264	29.660
38	1001001	265	29.670
39	1001010	266	29.680
40	1001011	267	29.690

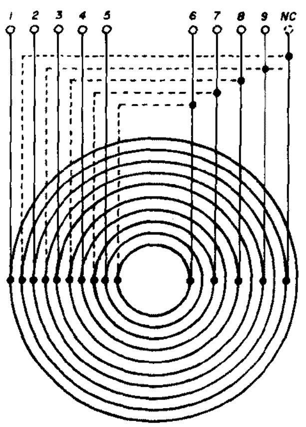


fig. 11. Internal finger contacts arrangement.

"track" of the PC board with a total of ten fingers. In the switch used, finger 10 had no external connection; that track was employed to make electrical connection to the outside track. The dashed lines show the effective position of the second set of fingers, 180 degrees from their true electrical position. It should be evident that, while encoding any particular division factor, it is necessary to alternate between the left and right side of the disk to assemble a digital word. This is a result of the interleaving contact arrangement used in these switches.

fm band plan

When considering how to "channelize" these boards, we tried for as rational an approach as possible. Initially we considered the original CB channel scheme. The channels are nominally spaced at 10 kHz, but there are several 20-kHz gaps, notably between channels 7 and 8, 11 and 12, 15 and 16, and 19 and 20. There was also a 30-kHz gap between channels 22 and 23. These gaps are presumably there to protect established users in the old pre-CB 11-meter band. When channels 24 through 40 were added, 24 and 25 were used to fill the gap between 22 and 23. The rest remained the same. Confusing, eh?

Since the switch used a printed circuit disk, it seemed possible to reprogram the switch to channelize our rigs as we desired. Thus we could eliminate the oddball frequency shifts one would experience when using a standard CB switch. The possibilities seemed endless, and numerous evenings were spent trying to establish a plan that seemed logical.

Since one cannot transmit fm on 29.7 MHz without having one's sidebands spilling out of the band, the top channel would have to be 29.69 MHz (channel 40). This fact seems to have been overlooked in the band schemes we have seen to date. If your rig can operate on 29.7 MHz, we recommend you not use that channel.

The first disk produced did not permit operation on the frequencies between 29.4 and 29.5 MHz. This was done to prevent interference to OSCAR Mode A downlink signals. This scheme made possible nineteen channels above 29.5 and twenty-one channels below 29.4 MHz. But 29.5 MHz, for some unknown (and apparently quite foolish) reason, is used as a calling channel. Transmitting on this frequency could cause severe interference to satellite beacon signals. We suggest a different channel be used as a calling frequency. Any suggestions?

Information available to us when the band plan was being worked out indicated that the present OSCAR 8 satellite was to be the last with a Mode A downlink. Since OSCAR 8 was then over three years old, it was reasonable to assume that three years hence it would be out of service. On that basis we fi-

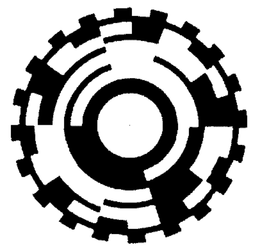


fig. 12. Disk PC layout.

nally decided on forty continuous, 10-kHz-spaced channels between 29.30 and 29.69 MHz. We simply would not use the channels in the satellite band until Mode A was no longer in use.

Then, long after the switches were designed and made, along came the new Russian satellites that not only use 29.4 to 29.5 MHz but frequencies between 29.3 and 29.4 MHz as well. Coordination began to seem impossible. All we can suggest is that prudence regarding use of transmitting frequencies be exercised.

Fig. 12 shows a 1:1 positive of our disk PC layout. The artwork should be 1.32 inches (33.5 mm) in diameter. It will be necessary to make a negative mask from this artwork. Be very precise when producing your disk, as the switch tracks are only 0.05 inch (1.25 mm) apart. Use 1/16-inch thick, 35-mm diameter, glass epoxy PC board and, if possible, tin plate the copper. The hole for the switch shaft can be made with a small file, using the original disk as a template.

in summary

These modifications and suggestions have all worked out quite nicely and helped several fellow hams get on the air sooner than might otherwise have been the case. See you on 10 fm!

acknowledgment

The idea of reprogramming the channel selector switch was proposed and developed by Ian Campbell, VE3IEO, to whom much credit is due.

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ham radio

design notes on a panoramic adaptor/ spectrum analyzer

Double-conversion superheterodyne with a 55-dB skirt filter, doubly balanced mixers, and a log detector

Human beings cannot see the radio signals that are everywhere around them. Hams spend much of their time listening to this or that signal, but their receivers let them hear only one at a time. Wouldn't a new dimension open up if you had a way of seeing those signals your radio wasn't tuned to — if you could see all the signals over a whole band of frequencies, rather than listen to just one of them?

A panoramic adaptor — a spectrum analyzer for engineers — can be built quite reasonably as a most useful accessory for the shack. Once connected to your receiver, the panoramic adaptor will give you rf vision.

A panoramic adaptor or spectrum analyzer will display the frequencies and magnitudes of all signals within some bandwidth (generally much wider than the bandwidth of your receiver) on an oscilloscope screen. For example, if your receiver is tuned to 14.200 MHz and the panoramic adaptor is set to scan plus or minus 50 kHz of your center-tuned frequency, all signals on 20 meters from 14.150 to 14.250 MHz will appear on the scope display. If someone on

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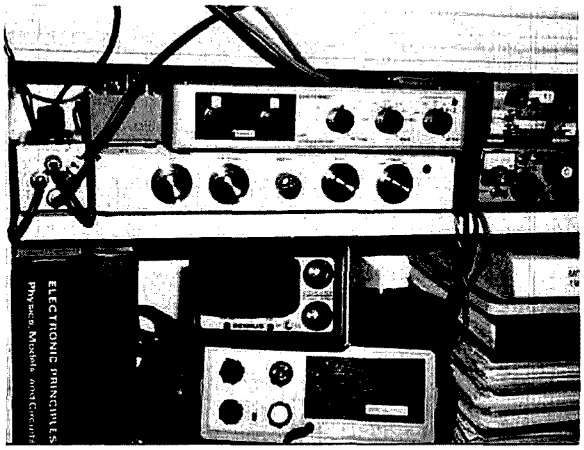
14.175 MHz gets on the air, his signal will suddenly appear as a pip on the screen, 25 kHz away (¼ the screen width), with a height proportional to the strength of his signal.



A view of the panadaptor. The white box is the actual unit with the display scope below; its power supply is on the right. A commercial general coverage receive-converter sits on top of the analyzer, with its power supply also at the right.

By adjusting the sweep-width for a plus or minus 10 kHz display, you can easily see dead spots on the band and plop your signal there for a schedule or CQ. Narrowing the sweep even further, you can analyze the modulation characteristics of the station you're receiving — such as upper or lower sideband, DSB, or a-m, fm, or even the shift or spacing between tones of an RTTY signal. For instance, a station running SSB with carrier injected — and telling his friends he's on a-m — can easily be distinguished from the full double-sideband a-m signals!

You can also see splattering, or readily identify the kilowatt station who's desensing your receiver's front end — he's the one up the band 30 kHz with the pip height almost off screen! You can get classical modulation patterns of a-m and fm signals, showing sidebands and odd order products. If you're a utility station listener (someone who likes to snoop on non-Amateur and non-broadcast high-frequency communications, like the strategic air command, coast guard, search and rescue, etc.), you can tune your receiver to an active band of frequencies, with the panoramic adaptor set for wideband scan, and zoom in on fleeting signals as soon as they pop up on the display. I've found dozens of hidden high-frequency signals normally missed when you're limited to the 2.1 kHz window of the basic receiver. With a VHF converter ahead of your high-frequency rig, the adaptor similarly lets you see and tune to those signals you'd usually miss if your receiver were just sitting at 50.110 MHz or 144.200 MHz.

history

The history of the panoramic adaptor, or spectrum analyzer, goes back to the 1930s and possibly earlier, when one could read in the *Proceedings of the IRE* (precursor to the IEEE) about various equipment designed to plot, on a crt or on paper, a magnitude versus frequency graph of the signals applied to its input. One such Fourier Analyzer (as they were sometimes called) had a motor-driven variable oscillator which slowly swept back and forth across its frequency range as the operator watched the output plot on the screen of a long-persistence cathode ray tube. Earlier models of spectrum analysis machines were actually mechanical devices devised to break a complex waveform into its Fourier (sine and cosine) components. They were full of gears and wheels. Some photos show them being operated by a hand crank.

Fortunately, modern-day spectrum analyzers don't need motors or even hand cranks, if you don't count an occasional knob-tweak as a cranking operation. In fact, the panoramic adaptor/spectrum analyzer to be described has some of the nicest modern devices at its heart: double-balanced mixers, wideband power amplifiers, a varicap diode-tuned oscillator, and an IC logarithmic detector.

basics of spectrum analysis

A spectrum analyzer is basically a narrowband filter swept through a band of frequencies with the resultant output plotted versus the frequencies you just swept through. Imagine that you had a tunable bandpass filter, and you tuned it, slowly, through the 20-meter band. As you proceeded up the band, the filter's output would increase every time you tuned through a signal, and then drop down when you went through an unoccupied part of the band. Now imagine that you hooked the output to the vertical plates of an oscilloscope, and, at the same time, you had the horizontal plates of the scope connected to the tuning knob of the filter, so that as you went up in frequency you'd move the spot from left to right. Now you've got a magnitude versus frequency display of 20 meters — a panoramic adaptor.

There are problems with this simple model. First, we want to be able to separate nearby stations, so the filter has to be very narrow. And it must be tunable, which makes it technologically almost impossible to build. Further, you would need a very high-

frequency scope to register the filter's output at 14 MHz; these are expensive, so we need to rectify the output and apply this dc signal to an inexpensive low-frequency scope. Finally, you don't want to sit and turn knobs all day to use the panadaptor; some kind of sweep generator is needed to do the work for you.

"... the spectrum analyzer shouldn't interfere with the receiver. This may seem obvious ..."

The solution is to build a superheterodyne spectrum analyzer, where we keep the filter fixed at some i-f frequency and, using a mixer, sweep the signals of interest through it with a scanning local oscillator. In fact, the model to be described is a double-conversion superhet, but that doesn't change the operational principle. As far as a detector or rectifier goes, an IC takes care of that in a single stage, with an excellent added feature to boot. The sweep generator is also fabricated from an IC, which drives a voltage-controlled local oscillator, nicely freeing the operator from strained wrists.

design goals

Getting down to the actual design goals of a modern panoramic adaptor: first, the device should have good dynamic range, displaying signals just above the noise floor of your receiver to those that nearly knock the S-meter off its pin. This makes a logarithmic detector a necessity, for it compresses a very wide linear voltage range (the 0.1 to 1000+ microvolts at your receiver's input) to a log scale that is easily viewed on one vertical scale on the oscilloscope. In addition, the analyzer itself should have wide dynamic range — not be susceptible to signal overloads. This design incorporates passive double-balanced mixers with fairly high-level injection and intercept points so these weak-links in the superhet circuit are practically overload-proof.

Next, the panoramic adaptor should have good resolution, the ability to separate signals from one another in a crowded band. This design incorporates a very narrow single crystal filter which is easy to build and which gives at least 55 dB skirts, and is inexpensive. Several options are available to those who want even better performance.

The adaptor should also have variable sweep width and rate, and some means to control its gain. It shouldn't respond to signals outside its bandwidth, and should give a linear frequency sweep out to about plus or minus 100 kHz. The display shouldn't cost a fortune; here the adaptor will work with any old clunk of an oscilloscope, as long as it has dc-coupled vertical and horizontal inputs. The scope's own sweep generator isn't even used; I bought my 3-inch display for $20 at a flea market and removed the sweep circuit tubes to save on heat generation. Vertical and horizontal amplifier frequency response is also unimportant — the adaptor essentially puts out dc.

Finally, the spectrum analyzer should not interfere with the receiver it's connected to. This may seem like an obvious requirement, but if you're interested in receiving signals from dc to 30 MHz, you don't want any local oscillator (LO) energy in that band of frequencies — any amount of LO leakage would be picked up by your receiver.

This panoramic adaptor was designed to work with the author's aging FT-101B which has a general coverage receive converter ahead of it. A simple modification of the LO and filtering will make this design work with any rig; all other components are broadbanded.

the circuit

The block diagram (**fig. 1**) shows the basic circuits used in this adaptor and gives figure numbers for each of the individual stages, **figs.** 2 through **13**. For sake of simplicity let's say the receiver to be used has an i-f of 3 MHz, and that you tap into it before the narrow receiver's filter, and that this wider-band part of the receiver i-f amplifier is about plus or minus 100 kHz wide.

The i-f signals go into a wideband amplifier with about 20 dB of gain, and then into the first double-balanced mixer (DBM). Here the signals are mixed with an amplified and filtered signal from the first local oscillator; this LO is voltage tuned and is connected to a sweep oscillator so the original 3-MHz signals are translated up to about 36 MHz. The image at 30 MHz is filtered, then the 36-MHz signals are amplified with another wideband amplifier with variable gain. Here another mixer is employed to beat the signals down to 5 MHz, using a crystal-controlled LO with amplifier and filter. At this 5-MHz i-f we have the narrow-crystal filter, which sets the resolution of the instrument, and feeds into a high-gain 5-MHz amplifier and log-detector IC. Power supplies and some op-amps around the sweep circuit complete the block description of the panoramic adaptor.

Everything is in modules or blocks which can be in-

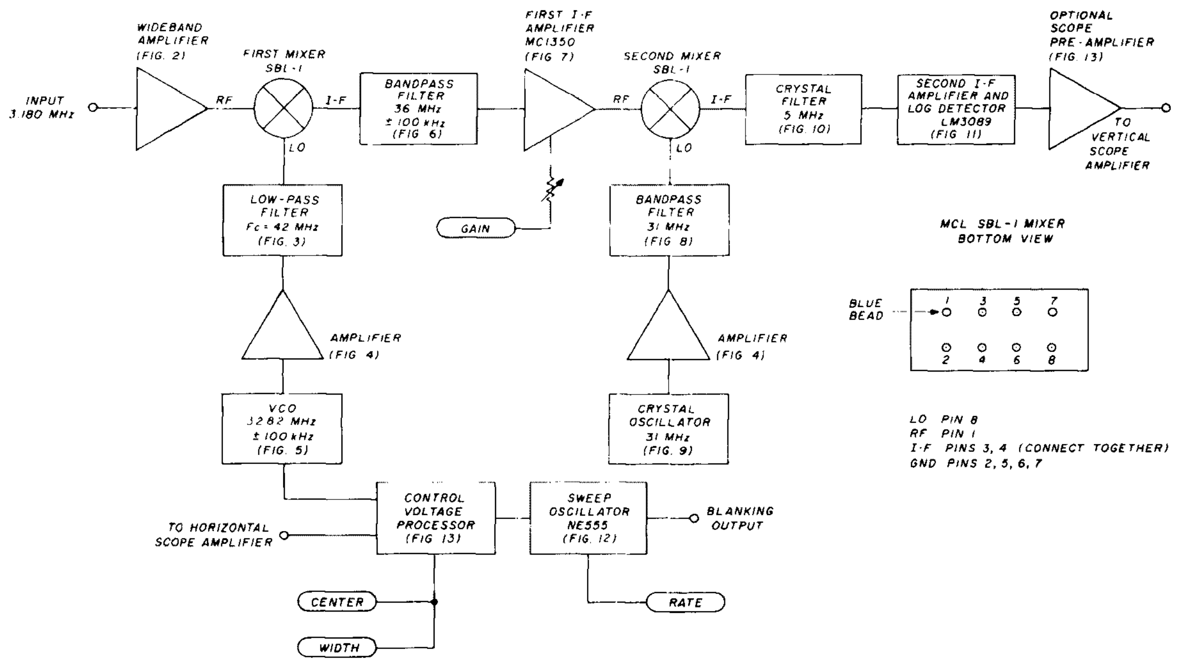


fig. 1. Block diagram of the panoramic adaptor. Interconnections are made with subminiature coax (RG-174/U or equivalent). The optional scope pre-amp is useful if the vertical sensitivity of your oscilloscope is marginal; it comes free as one of the op-amps in the LM324 control voltage processor.

dividually built and tested. A circuit of this complexity can't be thrown together all at once in a weekend; each sub-assembly needs check-out before it all goes together. In fact, the author's unit worked the first time around (save for one problem to be mentioned) once all the modules were built and operating individually. Of course, each module can be constructed and tested in a weekend; this spreads the project out and makes for an interesting diversity of circuits to explore over a couple of months.

The output from your receiver's i-f goes to a wideband preamplifier which provides some gain and essentially sets the noise figure of the panadaptor (the noise figure of the whole system is, of course, set by the front end of your receiver). Any wideband design will work satisfactorily here, since the inherent selectivity of the associated receiver's front-end and wideband i-f coupling will keep spurious signals from this amp. I used a commercially available amplifier; I provide a schematic of another design that works as well.

The signal now goes into the first mixer. Here you should use a passive double-balanced mixer, as mentioned above, for greatest dynamic range. The least expensive of these come from MCL (Mini-Circuits Lab, 2625 E. 14th St., Brooklyn, New York 11235) and will perform well; typical cost is $4-$5 each, which is what one vacuum-tube mixer would cost nowadays, anyway.

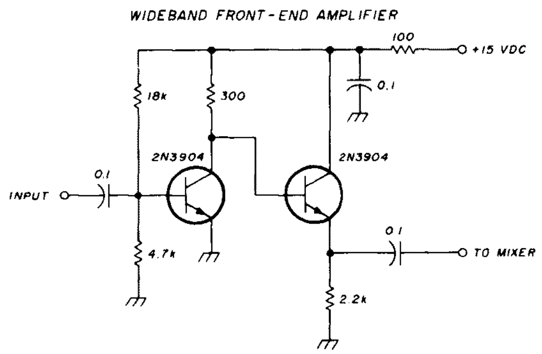


fig. 2. Wideband amplifier.

The i-f port of the DBM runs into a bandpass filter of conventional handbook design, scaled to the frequency of interest. Here, we want a center frequency of 36 MHz, with a passband flat over plus or minus 100 kHz of that frequency. The filter needs to reject the image at 30 MHz (29.64 ± 0.1 MHz) and this design provides some 40 dB of rejection. This topology was also checked on a computer analysis program and its performance is nearly as predicted. The 9 dB or so of passband loss it exhibits is mainly due to the finite Q of the inductors (which the computer can take into account), but this is not a serious problem since there's ample gain before and after this stage.

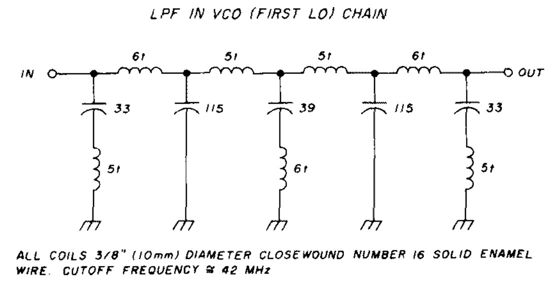


fig. 3. Low-pass filter, F_C = 42 MHz.

One of the most important design parameters of any superheterodyne receiver is LO purity; that is, you want the LO to put out just one signal and nothing else — no spurs, harmonics, or such. The first LO in the panadaptor is voltage tuned and free-running, using a varicap diode and FET as the active element. It's remarkably stable, within a kilohertz or so of its 33 MHz center frequency, and sweeps linearly over a plus or minus 100 kHz range. Following this LO is a power amplifier stage using broadband toroids and feedback; I modified the design from the *ARRL Handbook*. Here we get at least enough power to drive the first mixer, which requires +7 dBm input. But before we run this into the MCL mixer we filter it with a low-pass filter of conventional design. This filter was built without test equipment and later found to be non-critical as to exact components, as long as you're within a few percent of the values listed. Computer-aided analysis showed the theoretical and actual performance of the unit to be nearly identical. The cut-off frequency of this low-pass filter is about 42 MHz, and it does the job of rejecting all significant harmonics of the first LO and amplifier chain nicely.

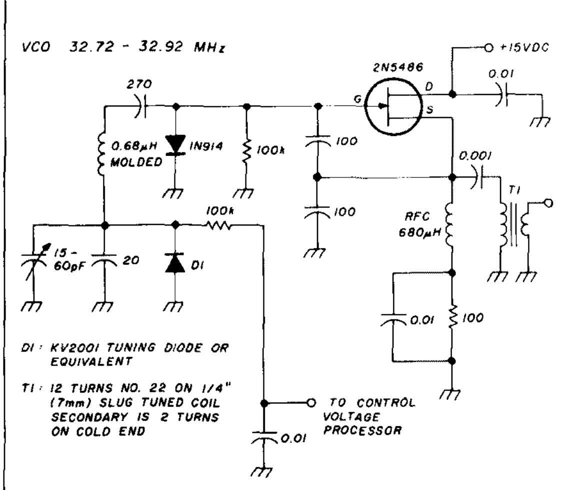


fig. 5. VCO, 32.82 MHz ± 100 kHz.

The filtered 36-MHz signals are now amplified by a wideband MC1350P i-f stage which is easy to build and get working. The input and output transformers can be homebrewed like the ones in the LO drivers, or a pair of MCL transformers can be used — or you can leave them out and only lose 3 dB or so of gain. This amplifier is capable of 36 dB of gain, and more importantly, can be cranked down to give about 30 dB of loss if necessary. Hence, it becomes the gain-controlled stage in the panadaptor, with a variable control on the front panel to change its gain as required.

These now-amplified signals run into the second mixer, another MCL DBM, which is driven by a crystal-controlled LO, another power-amp, and a narrowband filter. The LO and power amp are of conventional design, with the power amp identical to the one in the first LO chain. A 31-MHz crystal oscillator provides signals which, after amplification, go to a

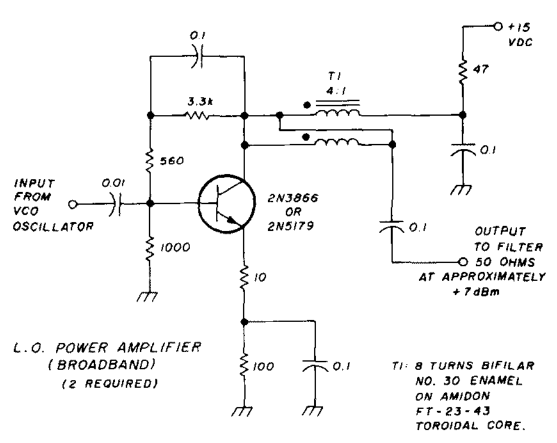


fig. 4. Amplifier.

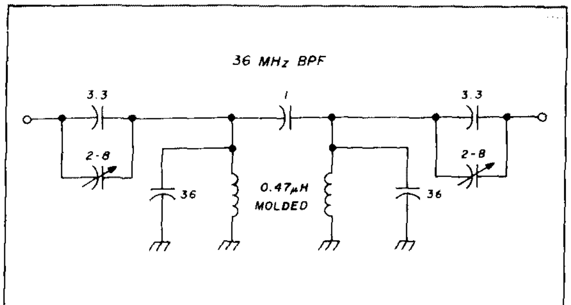


fig. 6. Bandpass filter, 36 MHz ± 100 kHz.

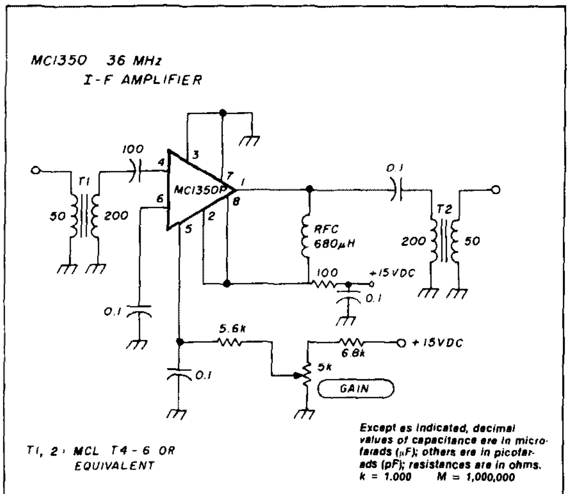


fig. 7. First i-f amplifier, MC1350P.

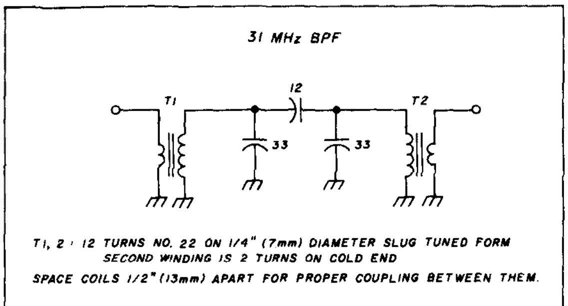


fig. 8. Bandpass filter, 31 MHz.

magnetically/capacitively coupled filter. This filter provides a very narrow passband to let only the 31 MHz signal through with about 1 dB of loss; it has a steep notch at approximately 41 MHz which nicely attenuates a spur there, and thereafter kills all other higher-order harmonics. The filter also has excellent return loss (good match to 50 ohms) at 31 MHz, which aids the stability of the preceeding amplifier.

The output of the second DBM, now at about 5 MHz, runs into a single crystal filter built from an article in *73 Magazine*. This unusual design has the crystal embedded in a two-transistor amplifier circuit, providing 15-dB gain at center frequency, with a 300-Hz bandwidth and skirts down to 55 dB below the peak. As mentioned above, this stage sets the resolution of the spectrum analyzer. Use a crystal with wire leads (any small unit will work — these are often available as surplus or microprocessor crystals) to minimize holder capacitance, and simply adjust the variable capacitor for equal skirt attenuation on either side of resonance. A sweep generator/scope arrangement is helpful for this adjustment.

Following the crystal filter is a surprising IC — the LM3089 fm i-f chip. This chip has a beautiful feature that makes it ideal for a panoramic detector. One of the pins is a tuning-meter output which, if you look at the spec sheet, gives an almost ideal logarithmic response to its input signals. This is the whole amplifier/detector circuit! There are no adjustments at all; the 5-MHz signal from the crystal filter goes in, gets amplified by the three i-f stages in the chip, then logarithmically detected. The rest of the IC (fm demodulator, muting, etc.) is not used and thus left unconnected. Though the chip is designed for 10.7 MHz service, it works fine at 5 MHz. At $3 a crack, the log detector feature (buried in its tuning-meter output) compares with commercial log-amps costing several hundreds of dollars.

The remaining circuitry is for the sweep generator and controls; here an ubiquitous NE555 timer IC plus a transistor makes for a very nice linear ramp generator, with rate variable from a few Hertz to a hundred Hertz or so. The generator is self-triggering, thus reducing circuit complexity from conventional designs which have two timers in series. Following this stage is a buffer (so circuit loading won't spoil ramp linearity), and two subtractors so the ramp signal to the VCO can be adjusted in amplitude about some adjustable dc value, thus giving you variable-sweep width and centering. The ramp is tapped off before these controls so it can be fed as a constant sweep source to the scope's horizontal plates.

Finally, the power supply is of straightforward design, using two three-terminal regulators and a full-wave center-tapped arrangement for the plus and minus 15 volts. The positive supply draws some 200 mA, while just a few mils are needed for the − 15 volt bus. A word of caution: the author had no trouble getting the whole adaptor working once each module was built and tested, except for the power sup-

ply. The analyzer worked fine, but a horrible wideband racket was heard in the receiver at certain frequencies every time the unit was powered up. At first I suspected the wideband amplifiers in the local oscillator chains, but a resistor connected across the 15-volt line produced the same receiver noise, with the adaptor completely disconnected! I then shunted the regulators right at their input and output pins with bypass capacitors (0.001, 0.1, and 33 μF). Thankfully, this cured the problem.

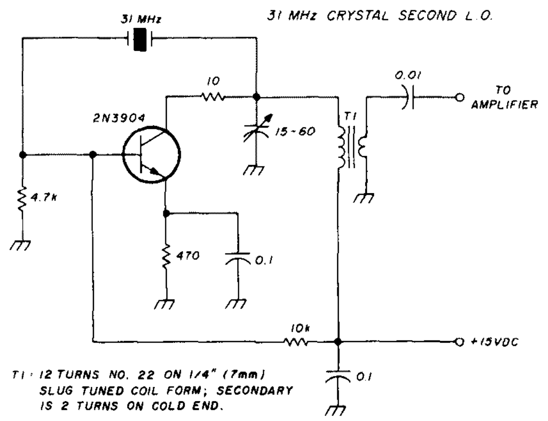


fig. 9. Crystal oscillator, 31 MHz.

a word on frequency scaling

Before I launch into some hints on building and adjusting each module, I should say something about adapting this design to other receiver i-f frequencies. Most receiver i-f's are below 9 MHz or so, and you can make the first and second local oscillators kick that up to some higher i-f in the analyzer, then back down to 5 MHz in its second i-f. In fact, most of the components are broadband at least up to 42 MHz (where the lowpass filter in the first LO cuts off), so no amplifiers need be redesigned. You will have to scale *the filter* components, *which should not be* difficult if you use a sweep generator/detector/scope arrangement to tune things up. There's nothing sacred about the 5-MHz crystal filter either; any fundamental crystal from about 4-8 MHz will work in the circuit, giving you even more flexibility. The important thing is to watch where you put your oscillator signals so they won't cause unwanted responses either in the analyzer or in your receiver.

For example, if you have a 9 MHz receiver i-f, use a

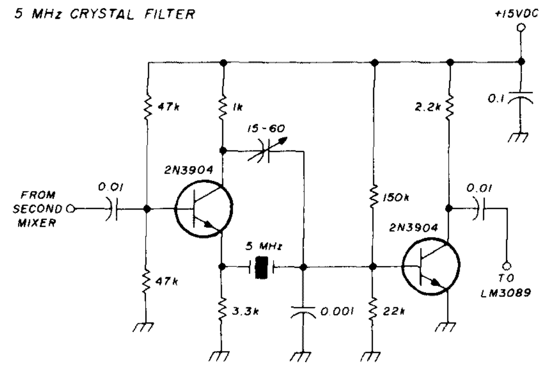


fig. 10. Crystal filter, 5 MHz.

voltage-tuned first LO at 32 MHz, filter the image and amplify at a first i-f of 41 MHz, then beat this down to 5 MHz with the second LO at 36 MHz. Here just the second LO filter, the 36 MHz bandpass filter, and a few tuned circuits in the oscillators need be scaled accordingly.

construction and tuning

My unit was built ugly style (no pc boards, just solder each component to a double-sided piece of copper-clad material), with small-diameter coax used to connect each module to one another. I even soldered the modules to a large piece of copper-clad board as a means of mounting them. The power supply was built in a separate box; a connector was used to lead power in and scope voltages out of the panadaptor chassis. Five controls on the front panel are gain, centering, width, rate, and power; the centering and

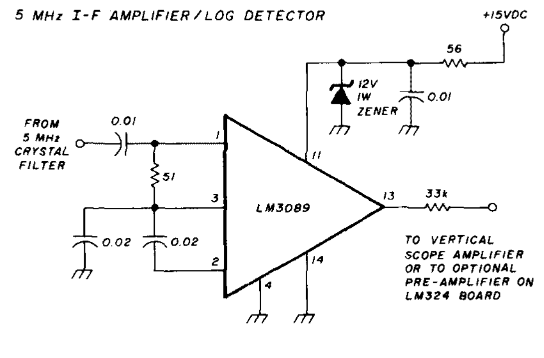


fig. 11. Second i-f amplifier and log detector, LM3089.

width controls are multi-turn for ease of use (the width control has a dial that can be calibrated if you wish).

To tune things up, note that only the filters and the LOs need be touched; all else is broadbanded. Tune each unit individually; when you hook all of them together you won't have to adjust anything.

For the first voltage-tuned LO, try to choose the LC network and the varicap diode so that several volts will swing the frequency over the desired 200 kHz of total spread. I padded down the tuned circuit until it did this — it's desirable because the LO won't be so sensitive to small pick-up voltages on the control line from hum or other sources. Using the values shown, I got the LO to cover 32.82 MHz plus or minus 100 kHz with a voltage swing from 5 to 9 volts. Don't make the swing too large as the ramp generator puts out a *maximum* of 5 volts peak-to-peak. The trimmer across the tuned circuit can be adjusted for center frequency (here 32.82 MHz to beat with the center of the FT-101B's i-f at 3180 kHz giving 36 MHz) with 7 volts on the tuning voltage input.

As mentioned above, the filters are best tuned with a sweep generator, scope, and detector. The 36-MHz bandpass is tweaked for flattest response over 36 MHz, and best rejection of 30 MHz. If you keep the frequency conversion scheme close to mine you'll not have to touch the 42-MHz low-pass filter in the first LO chain at all. The 31 MHz second LO filter is simply tweaked for maximum power out of the oscillator-amplifier stages, loaded into 50 ohms.

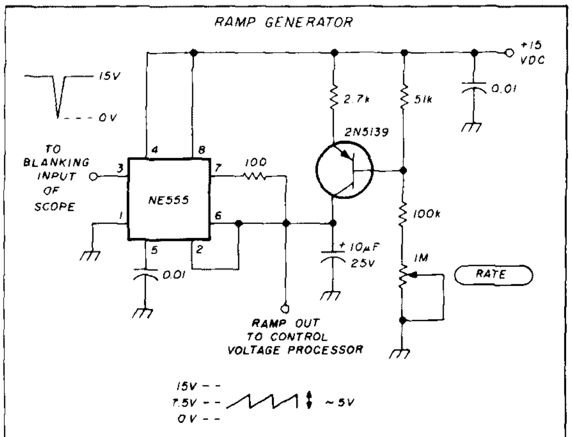


fig. 12. Sweep oscillator, NE555.

The two variable capacitors on the 31-MHz oscillator should be tuned for maximum power-out consistent with reliable starting. This LO, the power amp, and its narrow-band filter should be tuned as one unit.

As a check of the output power of the two LOs, you should use a vtvm with rf probe and measure the voltage across a 50-ohm resistor. For +7 dBm you should read 0.5 V rms, plus or minus 20 percent. You could also use the DBM as the load, as they are nominal 50-ohm devices and will be the actual load of the LO in use.

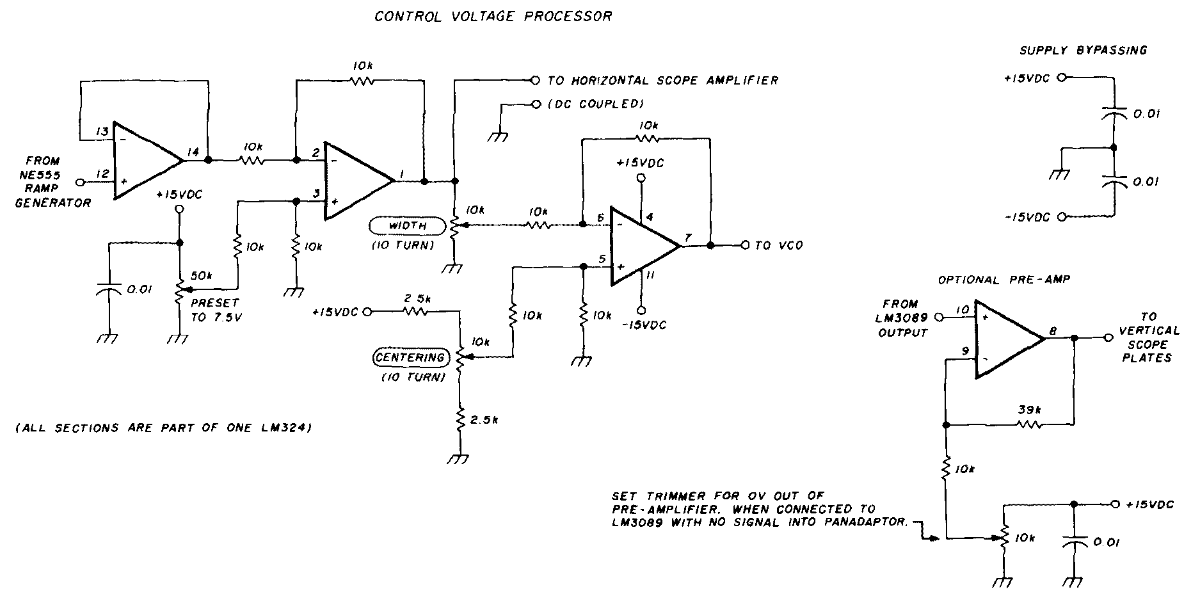


fig. 13. Control voltage processor and optional preamplifier.

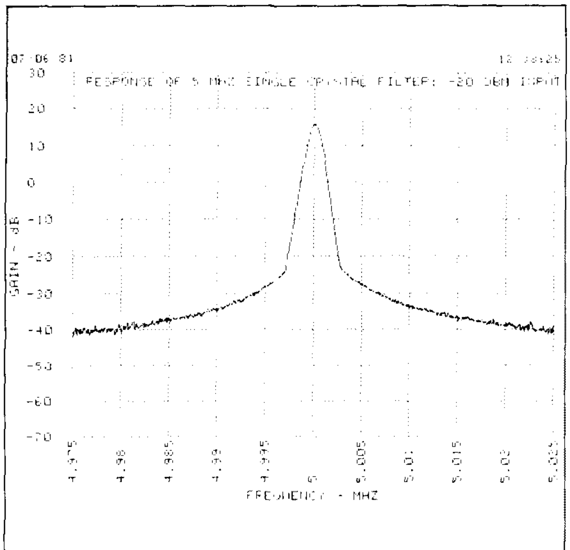


fig. 14. Measured bandpass characteristics of the 5-MHz crystal filter. Note ultimate attenuation is greater than 55 dB; the 3-dB bandwidth is less than 300 Hz.

The trimpot on the LM324 op-amp board should be set so that you get + 7.5 volts at the slider contact. This will help keep the pip centered on the screen as you decrease the sweep width for a closer look at some signal. It can be fine-adjusted after the unit is working as a whole.

Needless to say, you should always use short leads and bypass all power going to the various modules. The ugly method of construction makes it easy to use short connections since ground is all around on the board. My unit was assembled into a chassis with no shielding of any module from any other. It works fine this way, with no spurious pips on the display. However, another frequency conversion scheme may require some shielding, though the narrow range of the device (plus or minus 100 kHz) helps keep the possible spurs out of the tuning range.

hooking it up

If your receiver has an i-f output jack on the back, as many modern transceivers do, you merely run a shielded cable from that jack to the input of the pan-adaptor. If you notice that the unit is loading down the receiver, try a 4:1 transformer in the line, or put two in series to match the impedances.

Receivers without an i-f output jack can be connected by running a cable to the last i-f stage in the receiver before its narrow-band filter stages; this is usually after the receiver's last mixer. A small value coupling capacitor should be used at the tap-off point to avoid loading down or detuning the receiver's i-f amplifier.

The oscilloscope should be connected to the horizontal/vertical outputs of the analyzer; dc couple the scope's amplifiers and set them to approximately 1-volt/division. The blanking output of the adaptor sits at 15 volts and produces a narrow spike to ground upon retrace; my scope blanks almost completely with this input to its external blanking terminals. If you can't find your scope's blanking input it's not a big deal; the retrace is so fast compared to the forward sweep speed that you barely see the retrace under normal intensity settings anyway.

using the spectrum analyzer and some options

When you first turn things on, you've got the scope gain controls plus the four controls on the pandaptor to play with. Start with the gain high enough to see noise (grass in spectrum analyzer jargon) on the scope baseline, and adjust sweep rate and width to about center of rotation. Now turn on the 100 kHz calibrator in the receiver (or 1 MHz calibrator, or any strong locally-generated signal) and tune the receiver so you can hear this signal. Somewhere on the display you should see a large pip — if not, tune the

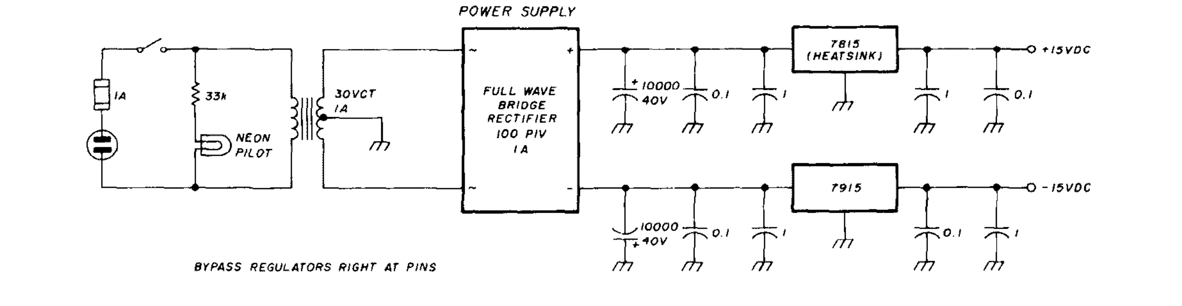
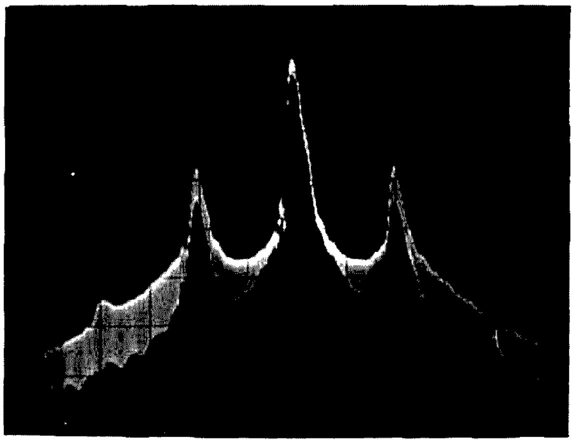
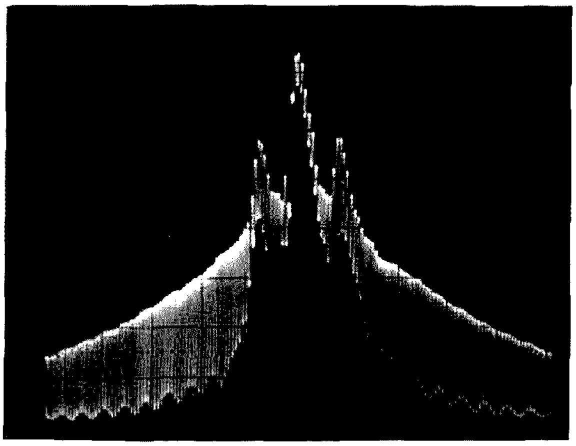


fig. 15. Power supply.

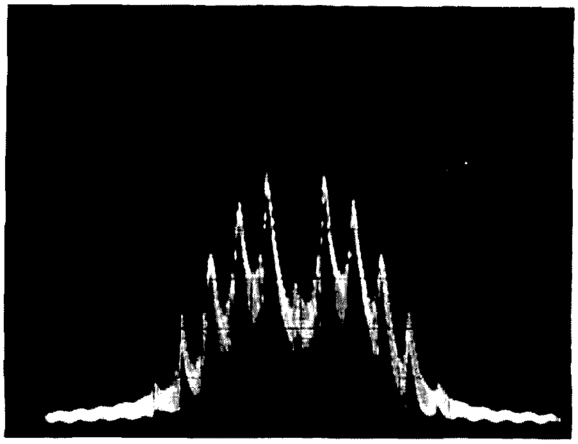
centering control until you do. Now set the gains on the adaptor and scope (the scope controls seldom need readjustment once set) for a presentation similar to those in the photographs. If you have a 25-kHz calibrator, flip that on and see the spectrum of signals displayed. In fact, this will calibrate the horizontal axis as you change the sweep width; you can set the center pip to the center, and the ones on either side to the edges of the display, for instance, to give you plus or minus 25 kHz of scan width. From here on, you will quickly learn what gain, width, and sweep speeds give the best display for a particular signal.



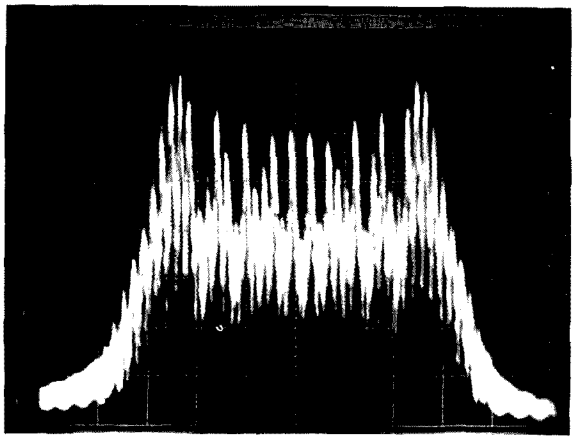
Panadaptor display with a 3180 kHz a-m signal at its input, 90 percent modulated at 1 kHz. Note the generator's intermod visible as another sideband set 2 kHz from the carrier.



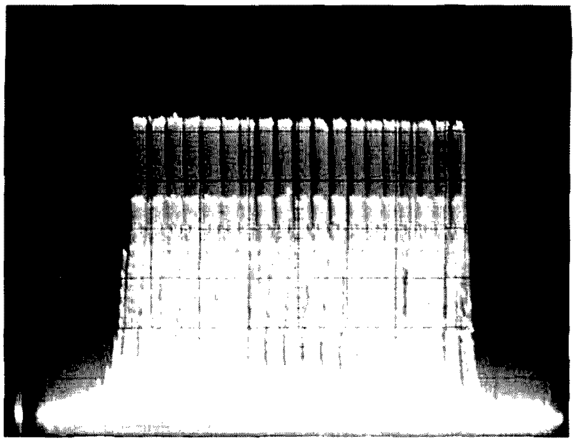
Same as photo 2, but with 400-Hz modulation. The sidebands are easily seen, demonstrating the instrument's 300 Hz or better resolution capability.



A classical fm waveform. Here at 2.3 kHz deviation with 1 kHz modulation. Note the carrier has almost disappeared, due to the Bessel-function character of fm for this particular modulation index.

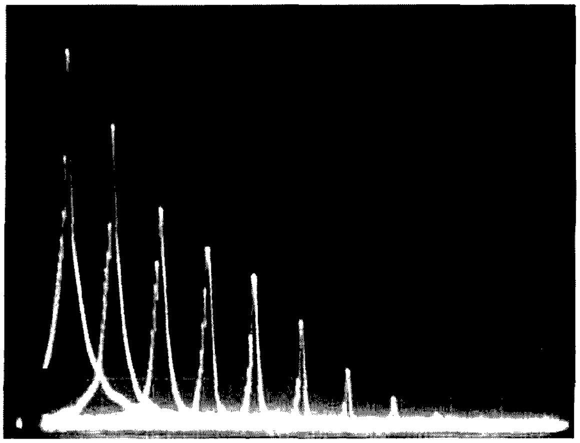


A wideband fm spectrum, 15 kHz deviation with 1 kHz modulation.



Sweeping ±50 kHz around the panadaptor's center frequency shows its excellent amplitude flatness over the band.

Note that if you increase the sweep rate too much, the pips get broader at the base and you lose resolution. Try to keep the rate slow enough to have good resolution yet fast enough so flicker is not bothersome. You may also notice that the resolution increases for narrower sweep widths, approaching 300 Hz as you get to very narrow sweeps. This is handy, for example, in checking the spacing of two tones in an RTTY signal, providing they're spaced further than 400 Hz or so.



A multiple-exposure photograph showing the analyzer's good frequency and amplitude linearity. Amplitude steps are in 5-dB increments and frequency steps in 5-kHz increments across the screen. Greater than 40 dB of range is displayed here.

No project is complete without some options; here you could get better than the 55-dB skirts of the simple crystal filter, by substituting a crystal lattice unit as used in commercial transceivers. Surplus Atlas filters are available for about $15 with center frequencies near 5 MHz; a narrow-band version of one of these would probably give very good performance. However, the band-edge roll-off of these lattice designs is very steep and could necessitate slow sweep speeds so the filter won't ring as signals move through them. Try it and see.

Another option would be a linear detector (as on a commercial spectrum analyzer), useful for some applications. Here the LM3089 comes in handy; it provides an amplified i-f output port that can be rectified in a linear detector circuit for later application to the scope's vertical channel. A switch could select between the log and linear displays.

This modern spectrum analyzer is an updated and improved version of the older panoramic adaptor, and besides the applications mentioned earlier in the article, will find other uses around the shack. With the rf vision it provides, a new facet of radio communication monitoring becomes possible, with rapid signal detection, modulation analysis, and band-condition assessment all easily accomplished. Soon you'll feel quite blind without its help, and you will switch it on every time you fire up the station receiver for a simple ragchew or just to snoop around the spectrum.

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ham radio

the Bragg-cell receiver

New uses for an old technology

The year is 1932 and radio as a technology is just beginning. A French physicist by the name of Louis-Marcell Brillouin is establishing grounds for what we today call the Bragg-Cell Receiver.

He notes the phenomenon of interaction between sound and light and proves that monocromatic light can be defracted in the presence of sound. The 1932 experiment involved a source of filtered light that illuminated a column of water into which sound was coupled. By properly adjusting the angle of incidence of the light source, the first-order diffraction line was observed to become more intense while the other lines were cancelled, presenting a graphic representation of this interaction. This angle of incidence was later called the Bragg angle, *and the* phenomenon became the basis for what we today call the Acousto-Optical (A/O) receiver, or, the Bragg-Cell Receiver.

The block diagram of a modern Bragg-Cell receiver is shown in **fig. 1**. It resembles a single-conversion receiver with a very wide i-f (40 MHz) centered at 70 MHz. Many rf signals are processed at the same time, as the passband of the front end is also 40 MHz wide.

Extreme care must be exercised in designing a system with a wide dynamic range, as there are no preselectors or narrow-band filters in this approach. The i-f amplifier is also a very high-dynamic-range power device providing about three watts of wide-band video signal to the Bragg-Cell transducer, which acts as a launcher or transmitter. A modern Bragg-Cell transducer and medium is a block of very pure crystalline material such as quartz or lithium niobate, which is approximately 1 cm × 1 cm × 1 cm. A pair of tuned electrodes are bonded to the side of this material. This is where the i-f signal is applied, as shown in **fig.** 1.

When excited with signals within the passband of the receiver, rf wavefronts are launched (or propagated) through the medium, changing its refractive characteristics accordingly (a form of spatial phase modulation). If a beam of monocromatic light is introduced at the Bragg angle, a panoramic display of all the signals within the i-f passband can be obtained on a projection screen. A helium-neon laser is used as the source of monocromatic light.

The deflection angle (the displacement from the center of the screen) viewed on the screen, and the intensity of the light spots, are directly proportional respectively to the frequency and the power of the

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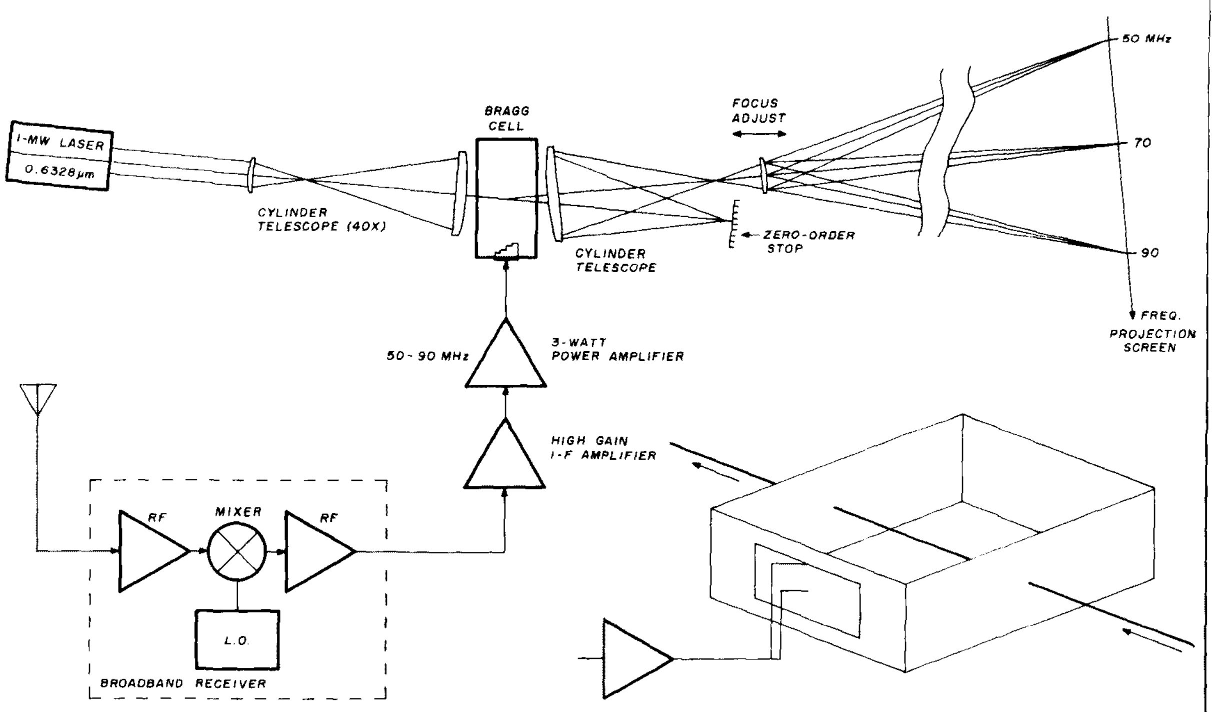

fig. 1. Block diagram of a modern Bragg-Cell receiver.

received signals. Post-detection demodulation can be accomplished by an array of closely spaced PIN photodiodes. The position of each photodiode corresponds to a specific frequency within the passband of the receiver's input, providing instantaneous reception of many signals, without sweeping or scanning.

The Bragg-Cell receiver can be viewed as a parallel processing device which converts radio-frequency energy to individual light spots positioned in the frequency domain on a dial-like base line.

While not completely understood from an application point of view, this receiver can be used as a wideband (non-sweeping) spectrum analyzer which can identify the presence of active frequencies. High-resolution programmable receivers can then analyze the particular signals.

The instantaneous nature of the Bragg-Cell receiver allows for a high probability of intercept (POI), since many signals can be observed at the same time. This, in turn, would make an ideal addition to a radio telescope which is searching many frequencies for extra-terrestrial signs of life.

The main disadvantages of the Bragg-Cell receiver are its limited dynamic range (typically 40 dB) and its frequency resolution, which is limited by the mechanical arrangement of the photo-detectors.

For readers interested in experimenting with such a receiver, helium-neon lasers are available today from a variety of sources, and Bragg-Cells can be purchased commercially from Intra-Action Corporation, 3719 Warren Avenue, Bellwood, Illinois 60104.

This article was adapted from the author's book, "Radio Communications Receivers," No. 1393, by TAB Books Inc., Blue Ridge Summit, Pennsylvania 17214: $13.95 soft-bound, $19.95 hard-bound.

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ham radio

technical forum

Welcome to the *ham radio* Technical Forum. The purpose of this new feature is to help you, the reader, find answers to your questions, and to give you a chance to answer the questions of your fellow Radio Amateurs. As a new feature, the Technical Forum will be shaped by the type and number of letters we receive from you. Do you have a question? *Send it in!*

Have you ever published a circuit, or do you know of a circuit, that can be used to test Zener diodes? The circuit should test the voltage rating of the Zeners. — Pete Hons, W3PKH.

Yes. Quite a few articles over the years have been published in the Amateur journals. Here are three:

"Low Voltage Zener Tester," *ham radio*, November, 1969, page 72: The circuit measures Zener voltages up to 10 volts and makes it possible to check voltages of unmarked and surplus devices.

"Two Methods of Testing Unknown Zener Diodes," *CQ*, August, 1972, page 38: The first circuit uses a 250-volt power supply, one fixed and one variable resistor, and any VOM or VTVM. It determines the breakdown voltage or a short or open condition. The second method feeds an audio signal to the Zener. The diode's characteristics are then read off a scope display.

"Bargain Zener Classifier," *73*, August, 1979, page 46: A method similar to that described above places an increasing voltage across the unknown Zener until breakdown occurs.

Has anyone designed an inductance meter that is fairly accurate at inductances of 20 μH on down to 0.01 μH? I built one that is used in conjunction with a digital volt-ohmmeter. However, its accuracy at 10 μH and down is very poor. — Gustave C. Budina, K9EBA.

When I first sat down to research your question, I thought finding the answer would be quite simple. But I found that most of what's been written has been devoted to L-measuring devices that go only as low as 20 to 50 μH. Below 20 μH very little has been done. One possible solution to your problem is discussed in the February, 1981, issue of *QST*. There, WA2TNG wrote of an inductance meter and frequency counter he designed that would measure values from 1 mH to 0.05 μH. When he started, he found that his design would work down to 1.4 μH. Below that, his inductance meter would not perform accurately. His design is basically a Colpitts oscillator fed into a frequency counter. Inductance is measured by determining resonant frequency and then calculating inductance using this formula:

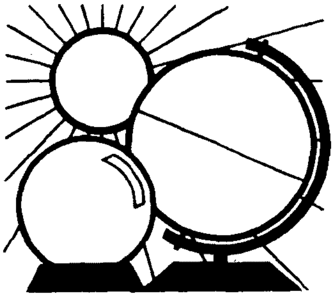
$$L_{\mu H} = (15.915/f_{MHz})^2 - L_o$$

L_o being derived from the Handbook LC chart for resonant frequencies above 5 MHz. Basically it is a correcting subtraction.

In looking into the reasons why the counter would not measure below 1.4 μH, the author found that the ceramic disc capacitors he was using had too much internal resistance (in excess of 7 ohms) and that this resulted in excessive loss in the tank circuit. Replacement of all capacitors with those of polystyrene design (internal resistance of 1 ohm) reduced series resistance. This resulted in an ability to measure inductance to values as low as 0.05 μH.

My suggestion is to look closely at your capacitors and replace them with the polystyrene type. That should give your inductance meter a greater range and give you a much more useful instrument.

ham radio



DX FORECASTER

Garth Stonehocker, K0RYW

last-minute forecast

Excellent DX conditions on the higher frequency bands (10-20 meters) during the third week and on into the beginning of the fourth week of February are expected, after a slow beginning the first two weeks. In fact, during these first two weeks of the month it may be better to try nighttime DX on the lower frequency bands. Those bands should be very good throughout the month unless disturbances prove particularly severe, which may be the case on the 1st, 11th, 19th, or 28th. Disturbances on the 1st and 28th should be less severe but four to six days in length.

No significant meteor showers occur during February. The full moon is on the 27th and lunar perigee the 25th. February is often the month of the highest mean solar flux values of the year, which results in very high maximum usable frequencies (MUFs). The ionization responsible for these MUFs moves toward the sides of the geomagnetic equator, giving trans-equatorial (TE) one-long-hop propagation. Geomagnetic disturbances, however, can enhance TE and lower mid-latitude MUFs. These disturbances result from particles in the solar wind traveling to earth and spiraling down into the polar regions. High solar winds result from solar flares and thin spots in the sun's corona.

Geomagnetic disturbances associated with thinning of the solar corona (coronal holes) should be increasingly affecting radio propagation. The reason is that the solar flux has rounded off the eleven-year peak and it has started its maximum rate of decrease; also, with decreasing flare output, the solar pressure on the earth's magnetosphere has decreased. This leaves the geomagnetic field sensitive to the solar wind (particles radiating from the sun) streaming through coronal holes. Geomagnetic disturbances from these holes are weaker (A of 20-30), longer (four to six days), and build up gradually. This is in contrast to disturbances following flares, which are intense (A often above 50-60), short (two to three days), and start suddenly. Coronal hole disturbances tend to recur in twenty-seven-day cycles similar to those of flares, but they tend to occur around the twenty-seven-day solar minimum rather than the time of maximum flare activity and flux.

Coronal hole disturbances reoccur so regularly that they have been labeled recurrent geomagnetic disturbances. In certain years they can be observed for as many as four to six consecutive twenty-seven-day cycles, and with as many as two to three distinct groups within a twenty-seven-day period. That's a lot of disturbance. Expect these recurrent disturbances to become more frequent soon and last through most of 1986 until the solar cycle, at minimum flux in 1986/7, begins to turn upward. The maximum disturbance is expected in the later part of 1984.

What does all this mean in terms of propagation and DXing? Well, simultaneous with the decreases in MUF toward sunspot minimum (see October, 1982, "DX Forecaster") will be 2½ years of increase in long-duration (four to six day) disturbances. When disturbed, the ionosphere becomes depleted south of the auroral zone (60° to 70° north or south latitude) in a region known as the trough. Where has the ionization gone? It has diffused up the geomagnetic lines of force to the geomagnetic equator. The more and longer the disturbance, the wider the trough becomes. This affects east/west paths from mid-latitudes (U.S.-Europe) by lowering the MUF while at the same time raising the MUF for TE paths.

band-by-band summary

Ten and fifteen meters will be open for worldwide DX from sunrise until after sunset during the twenty-seven-day solar flux maximums. Skip of 2500 miles (or multiples) is possible, and will follow the sun across the earth.

Twenty meters will be open to some *area of the world for the entire* twenty-four-hour period on many days of the month. The band should peak in all directions just after local sunrise, and again toward the east and south during late evening hours. During hours of darkness the band will peak toward the west in an arc from southwest through northwest, encompassing Pacific areas.

Forty and eighty meters will be the most usable nighttime DX bands. Most areas of the world will be workable from dusk until sunrise. Hops shorten on these bands to about 2000 miles for 40 meters and 1500 miles for 80 meters, but the number of hops can increase because signal absorption in the ionosphere's D-region is low during the night. The path follows the direction of darkness across the earth, similar to the way in which the higher bands follow the sun.

One-sixty meters will be similar to 80 meters, providing good working conditions for enthusiastic DXers who like to work the nighttime and early morning hours, especially at local dawn.

ham radio

WESTERN USA

GMT	PST	N ↑ ASIA FAR EAST	NE ↗ EUROPE	E → S. AFRICA	SE ↘ S. AMERICA	S ↓ ANTARCTICA	SW ↙ NEW ZEALAND	W ← OCEANIA AUSTRALIA	NW ↖ JAPAN
0000	4:00	10	20	15	10	20	10	10	10
0100	5:00	15	20	15	15	15	10	10	10
0200	6:00	15	20	15	15	15	10	10	15
0300	7:00	20	40	20	15	15	15	10	15
0400	8:00	20	40	20	20	15	15	15	15
0500	9:00	—	40	20	20	15	15	15	20
0600	10:00	—	—	20	20	15	20	15	20
0700	11:00	—	—	20	20	20	20	20	40
0800	12:00	—	—	20	20	20	20	20	40
0900	1:00	—	—	20	20	20	20	20	40
1000	2:00	—	—	—	20	40	40	—	40
1100	3:00	—	—	—	—	40	40	—	—
1200	4:00	—	—	—	—	40	40	—	—
1300	5:00	—	—	—	—	—	40	—	—
1400	6:00	—	20	20	—	—	—	—	—
1500	7:00	—	15	15	15	—	—	—	—
1800	8:00	20	15	10	10	—	—	20	—
1700	9:00	20	15	10	10	—	20	15	—
1800	10:00	20	15	10	10	—	15	15	—
1900	11:00	15	15	10	10	15	10	15	—
2000	12:00	15	20	15	10	15	10	10	20
2100	1:00	15	20	15	10	15	10	10	15
2200	2:00	15	20	15	10	15	10	10	15*
2300	3:00	15*	20	15	10	15	10	10	10

FEBRUARY

MID USA

MST	N ↑ ASIA FAR EAST	NE ↗ EUROPE	E → S. AFRICA	SE ↘ S. AMERICA	S ↓ ANTARCTICA	SW ↙ NEW ZEALAND	W ← OCEANIA AUSTRALIA	NW ↖ JAPAN	CST
5:00	15	20	15	10	15	10	10	10	6:00
6:00	15	20	15	15	15	10	10	15	7:00
7:00	15	40	20	15	15	10	15	15	8:00
8:00	20	40	20	15	15	15	15	15	9:00
9:00	20	40	20	20	15	15	15	20	10:00
10:00	—	40	20	20	15	15	20	20	11:00
11:00	—	40	20	20	20	20	20	40	12:00
12:00	—	40	20	20	20	20	20	40	1:00
1:00	—	—	20	20	20	40	20	40	2:00
2:00	—	—	—	20	20	40	20	40	3:00
3:00	—	—	—	—	20	40	—	—	4:00
4:00	—	—	—	—	40	—	—	—	5:00
5:00	—	—	—	—	—	—	—	—	6:00
6:00	—	—	15	15	—	—	—	—	7:00
7:00	20	20	10	10	—	—	—	—	8:00
8:00	20	15	10	10	—	—	—	—	9:00
9:00	20	15	10	10	—	—	20	—	10:00
10:00	20	15	10	10	—	20	15	—	11:00
11:00	20	15	10	10	20	15	15	—	12:00
12:00	20	15	10	10	15	10	15	—	1:00
1:00	20	15	10	10	15	10	10	20	2:00
2:00	—	20*	10	10	15	10	10	15	3:00
3:00	—	20	10	10	15	10	10	15	4:00
4:00	—	20	15*	10	15	10	10	10	5:00

EASTERN USA

EST	N ↑ ASIA FAR EAST	NE ↗ EUROPE	E → S. AFRICA	SE ↘ CARIBBEAN S. AMERICA	S ↓ ANTARCTICA	SW ↙ NEW ZEALAND	W ← OCEANIA AUSTRALIA	NW ↖ JAPAN
7:00	15	20	15	10	15	10	10	15
8:00	15	20	15	15	15	15	10	15
9:00	15	40	20	15	15	15	15	20
10:00	20	40	20	20	20	15	15	20
11:00	—	40	15	20	20	20	20	20
12:00	—	40	20*	20	20	20	20	20
1:00	—	40	20	20	20	20	20	40
2:00	—	40	—	20	20	20	40	40
3:00	—	40	—	20	20	20	—	40
4:00	—	—	—	40	20	—	—	—
5:00	—	—	—	40	—	—	—	—
6:00	—	—	20	40	—	—	—	—
7:00	—	—	15	15	—	—	—	—
8:00	20	—	10	15	—	—	—	—
9:00	20	—	10	15	—	—	—	—
10:00	20	10	10	15*	—	—	15	—
11:00	15	10	10	15*	20	—	15	—
12:00	15	10	10	10	15	20	15	—
1:00	15	15	10	10	15	15	15	—
2:00	—	15	10	10	15	15	15	—
3:00	—	15	10	10	15	15	10	—
4:00	—	20	10	10	15	10	10	20
5:00	—	20	15	10	15	10	10	15
6:00	15	20	15	10	15	10	10	10

*Look at next **higher** band for possible openings.

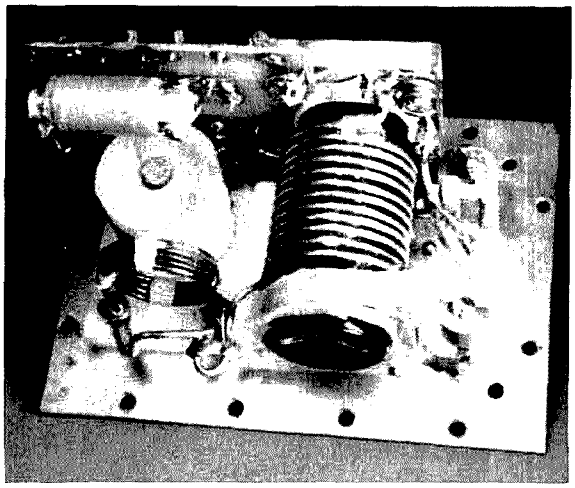
VFOs tuned
by cylinder and disc

Inductive tuning provides tailor-made frequency coverage and temperature stability

This article describes further results in my search for simple, reliable, variable-frequency oscillators (VFOs) reported in the July, 1980, issue of *ham radio*.[1] I feel there is a need for low-current-drain, stable VFOs of simple construction, especially for portable use. Also, the alternative of inductance-tuning techniques, rather than reliance on increasingly scarce, bulky, precision variable capacitors, offers great advantage.

Although the VFOs described here are simple, one could control them from a synthesizer output.[2]

When the VFO is to be used in communications equipment, one must avoid generating unwanted frequencies, which would have to be filtered out. This can be a problem with synthesizers. For Amateur use, simplicity may be the best approach. As a



An example of a cylinder-tuned VFO.

By **Richard Silberstein, WØYBF,** 3915 Pleasant Ridge Rd., Boulder, Colorado 80301

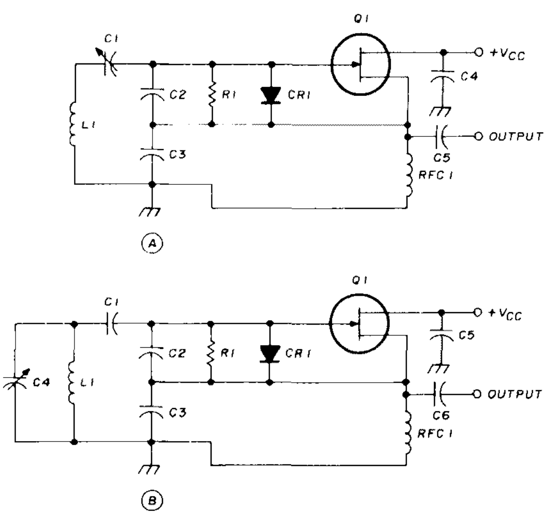


fig. 1. The well-known series-Colpitts oscillator is shown at A. C1, for tuning, can be placed at the other end of the coil. A basic Seiler oscillator, B, is adapted from the series-Colpitts. C1 is now a coupling capacitor and C4 is generally used for tuning. If L1 is tuned, C4 can be used for frequency calibration.

home experimenter, I build only analog frequency sources, and use only one frequency conversion in receivers. I plan to have the local oscillator on the high side of the received frequency, where possible, so as to place image frequencies and harmonics as far out of reach as possible.

This text begins by presenting the next development after those covered in reference 1: a Seiler oscillator in which incremental inductance tuning is done by means of a metal-foil triangle on a rotatable insulating cylinder adjacent to a coil.

The next development is a successful venture into frequencies higher than those usually used in a simple VFO. A 30-MHz VFO is tuned by means of a disc with a metal pattern, rotated close to the ground end of a high-*Q* inductor. Here, I returned to the Hartley circuit; a modification dispelled my earlier objections and left me with a very good, simple oscillator circuit.

Finally, I discuss temperature effects and rectify some wrong guesses made in the earlier article.

the cylinder-tuned oscillator

The Seiler circuit used in this oscillator is an outgrowth of the well-known series-Colpitts circuit illustrated in **fig. 1A**. C1 is conventionally the tuning capacitor, and C2 and C3 provide feedback. The value of frequency is determined essentially by L1 and a capacitance which has the value of C1, C2, and C3 in series.

In the interest of stability, it is desirable to have a high-*Q*, stable, frequency-determining circuit isolated as much as possible from the semiconductors. The Seiler circuit, **fig. 1B**, is an attempt to provide these conditions. Here, C2 and C3 are still the feedback capacitors, but C1 is now a coupling capacitor. There is a new capacitor, C4, which, in parallel with L1, would determine the oscillator frequency if C1 were small enough. In a practical circuit, if C1 is too small the drain current becomes excessive, and a still smaller value of C1 causes normal oscillation to cease. However, the advantage of the circuit to the experimenter is that he can adjust C1 to a practical, limiting value for at least partial isolation of L1C4. The higher the *Q* of this *LC* combination, the greater the possible isolation.

In the usual Seiler oscillator, C4 of **fig. 1B** is used for tuning. In my inductance-tuned version, useful only for a relatively small band of frequencies, there is a small pickup coil in series with the ground end of L1. The coil is coupled to a rotatable acrylic cylinder around which is glued a metal-foil triangle. This is shown in **fig. 2**. Because the magnetic field of the pick-up coil produces eddy currents in the foil, an opposing magnetic field is generated which reduces the inductance of that coil. This effect increases as the area of metal coupled to the coil increases, up to an area a little greater than that encompassed by the coil cross-section. Thus, as the larger end of the triangle is rotated into the field of the coil, inductance decreases and the frequency goes up.

Other designs may produce the same effect by using a fixed-metal surface or object and varying its distance from the concentrated magnetic field of the coil, or by varying its angle, or both. The copper-vane tuning element is one form of such a tuning device where a sickle-shaped rod made of a tube or wire is rotated into a half-toroid coil.[1] The cylinder and the disc, to be described later, have two advantages over the copper vane:

1. The useful tuning range is at least 270 degrees.

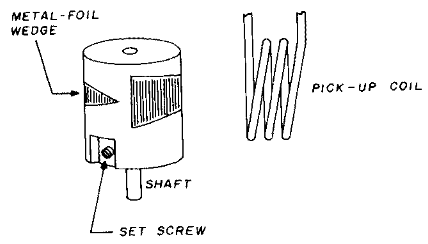


fig. 2. The cylindrical acrylic-spool tuning element.

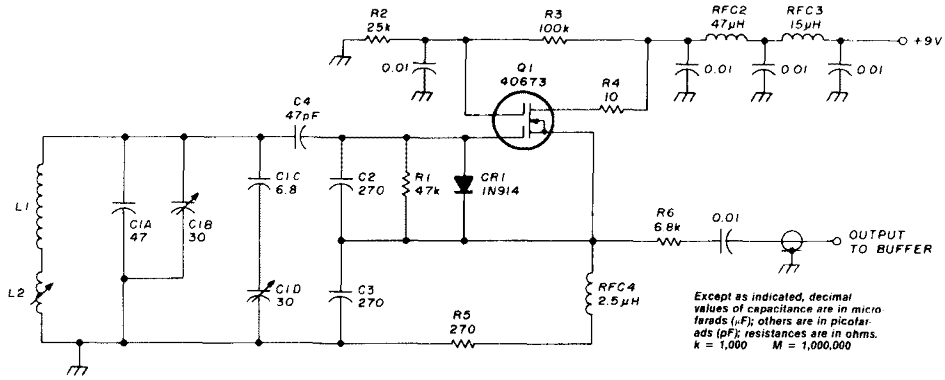


fig. 3. A 9-MHz cylinder-tuned VFO. Variable capacitors are air trimmers; C1B is used for coarse calibration and C1D for fine calibration. Tuning is done in the manner of *fig. 2.* L1 = 11-1/4 turns No. 16 tinned wire on a glass vial 1.04 inches (26.4 mm) in diameter and 1-1/4 inches (28.6 mm) long. L2 = 3-3/4 turns No. 16 enameled wire, close wound, 7/8 inch (19.4 mm) in diameter.

2. No detent (stop) is needed; the dial can be rotated continuously.

Fig. 3 shows a practical form of a cylinder-tuned Seiler for 9 MHz. Because the cylinder provides only the variable part of the tuning, it is necessary to provide means for setting a calibration point against a standard. C1B provides coarse initial calibration, and C1D, in series with a small capacitor, C1C, provides fine adjustment. This last function could also be performed by electronic means.

Fixed frequency-determining capacitors in this oscillator were originally polystyrene capacitors. The adjustable capacitors are air-dielectric trimmers.

The acrylic cylinder is 1 inch (25.4 mm) in diameter. The metal-foil triangle is made of 0.006 inch (0.15 mm) thick copper, 1 inch (25.4 mm) wide at the widest point. Heavy-duty household aluminum foil works almost as well as the copper.

I built the main inductor, L1, as an air-core solenoid with the same inductance as the 38-1/2 turn air-core toroid described in reference 1, since I believed at that time that the toroid was a source of positive frequency drift with increasing temperature. The coil consists of 11-1/4 turns of No. 14 tinned wire spaced on a glass (not plastic) medicine vial 1.04 inches (26.4 mm) in diameter for a total length of 1-1/8 inches (28.6 mm).

To make the coil, I first wound the wire on a dowel suspended between a lathe chuck and a "steady rest." This can be done manually, if you have help. I got just the right diameter of dowel by applying masking tape to a dowel which was a little too small. The coil, which slid off the dowel, was just a bit narrower than the glass vial but could be sprung and slipped onto the bottle so the turns could be spaced and would exert a clinching force. I had previously made end mounts by using a fly cutter on square pieces of acrylic to make holes the same diameter as the bottle. I placed terminals and mounting brackets at the corners and cemented the mounts to the bottle. Finally I anchored the coil to the vial surface by laying on strips of epoxy glue. I avoided coating the whole coil because of the tendency of most adhesives to absorb moisture or otherwise lower the *Q* of the coil.

The *Q* of this coil measured at 12 MHz was approximately 290. Partly because of the high *Q*, it was possible to have an oscillator current drain of only 1.8 milliamperes.

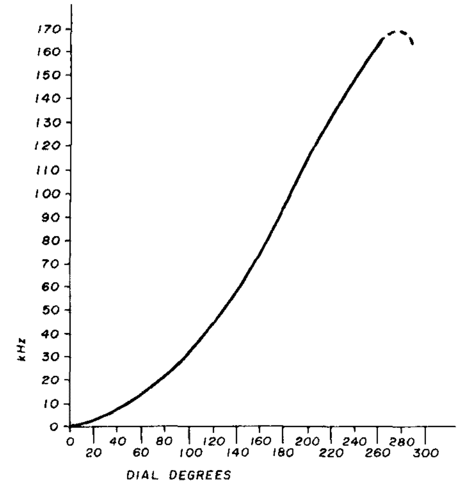
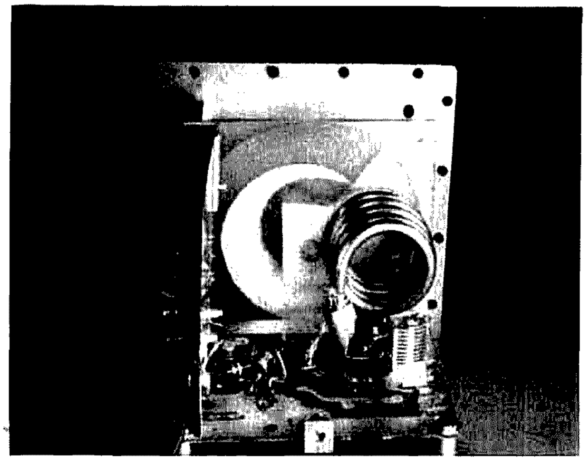


fig. 4. Incremental frequency calibration of the 9-MHz cylinder-tuned VFO.



A disc-tuned VFO.

RFC4 in the source circuit is a 4-section 2.5-mH air-core choke, used to minimize temperature-change effects on the oscillator frequency, which might occur with a simpler choke. The 270-ohm resistor, R5, isolates the choke, and may also prevent low-frequency oscillations. Resistor R6, 6800 ohms, helps to isolate the output cable and other output-circuit elements which might influence the oscillator frequency.

Fig. 4 is a plot of frequency change against angular shaft rotation for the 9-MHz cylinder-tuned oscillator. The useful tuning range is seen to be about 165 kHz, reasonably linear over most of its coverage.

the disc-tuned oscillator

Development of this oscillator coincides with a return to the Hartley configuration, as a result of a simple modification which eliminated some problems. Of course, any inductance-tuning method will work with any oscillator in which an inductance is a frequency-determining element.

In the simple Hartley oscillator of **fig. 5A**, feedback is produced in the coil. The alternating FET drain current through turns B-O on L1 induces a voltage on turns O-A, exciting the FET gate in the correct phase to sustain oscillations. The frequency is largely determined by L1C1, but instability can be introduced by the attached circuitry. The tapped-coil Hartley of **fig. 5B** is an attempt to isolate L1C1 from relatively unstable parameters. Tapping down the coil performs the same isolation function as adjusting C1 in the Seiler oscillator of **fig. 1B**.

Previous experiments had demonstrated a tendency of the circuit to break into parasitic oscillations as the taps A and O were moved down the coil. At the higher frequencies the turns between A and A′ act as a choke, isolating C1. Then the turns A-O-B, plus incidental capacitances and inductances of the etched conductors (or wiring), plus what is contained in all the attached circuit elements, are probably what produce these VHF parasitic oscillations. In **fig. 5B**, the area of greatest concern is enclosed by a dashed line.

In the modified Hartley of **fig. 5C**, the inserted resistor R2, of the order of 10 ohms, breaks up the loop which might resonate at a parasitic frequency. The oscillator I built using this circuit produced the desired results. Now, the rf choke in the FET source circuits of **fig. 1**, (RFC4 of **fig. 3**), was no longer needed and the feedback capacitors C2 and C3 were also eliminated. Relative isolation of the frequency-determining elements could be increased by moving taps A and O down the coil (until drain current became excessive).

In the usual Hartley oscillator, tuning is done by capacitor C1. In this oscillator, inductance tuning is accomplished by means of a rotatable insulated disc bearing a metal plate of such configuration that the ground end of a coil coupled to the disc is exposed to a varying area of metal as the disc is rotated. The

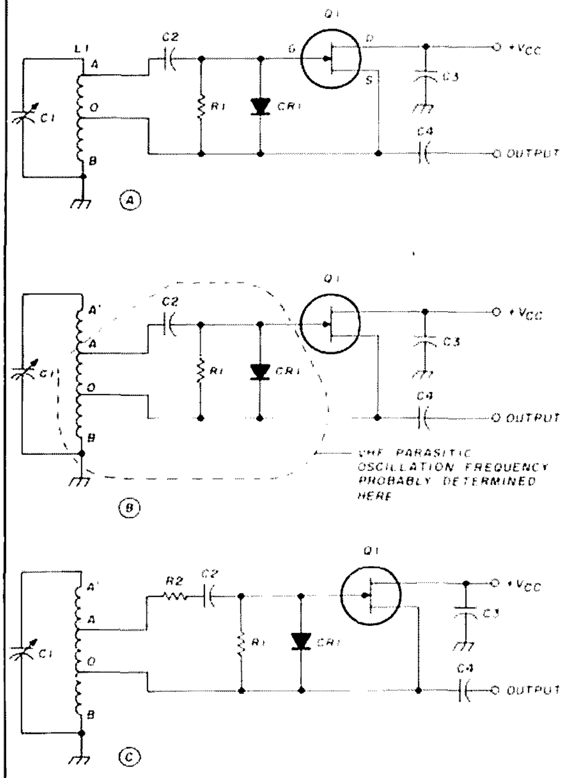


fig. 5. A simple Hartley oscillator (A); a tapped-down Hartley, (B), showing probable source of parasitics; introduction of resistor R2 in (C) breaks up parasitic resonances in the tapped-down Hartley.

simple geometric design I used was a triangle similar to that of the cylindrical tuner, except that I transposed the triangle to the polar coordinates of the disc. **Figs. 6A** and **6B** illustrate how I laid out the disc pattern.

I made the disc 3 inches (76.2 mm) in diameter, using single-side plated Fiberglas® epoxy-resin circuit board upon which I etched the pattern. Note that the pattern as transposed to the disc in **fig. 6B** is centered toward the outside of the disc, since the center of the pattern must rotate past the center axis of the coil, and the coil must be positioned so that one edge must clear the shaft and its support bearing. Refer to **fig. 8** for details.

To fasten the disc to its shaft, I made a simple, square acrylic hub. At the expected velocities of hand rotation, a dynamically balanced hub was not needed. I drilled and tapped two edges of the hub for set screws, and cemented it to the disc. To ensure a snug fit, I glued the hub to the disc and then drilled the final shaft hole after everything was cemented. To avoid heat flow of the plastic, and distortion of the hole, drill at slow speed, perhaps with the aid of a coolant such as cutting oil.

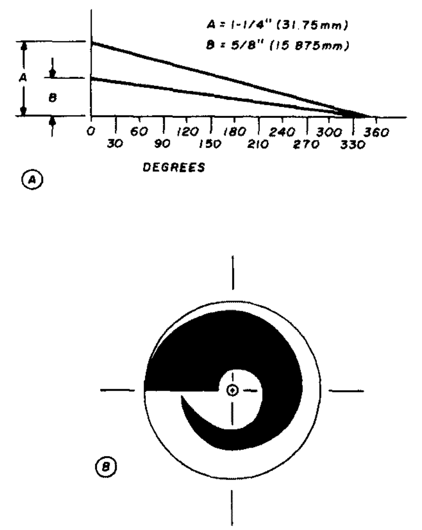


fig. 6. Scaling diagram for generating the tuning-disc pattern, (A); the pattern as etched on the disc, (B).

Fig. 7 shows the circuit of a modified, tapped-down 30-MHz Hartley VFO with buffer for 15-meter operation using a 9-MHz i-f.

The inductor, L1, is an air-core solenoid wound with 4-3/4 turns of No. 6 copper wire, 1-1/4 inches (32 mm) in outside diameter and the same length, with a Q of 260 measured near 25 MHz. The gate tap A is 1-3/4 turns above ground, and the source tap O is 3/4 turn above ground, which is at B.

Capacitor C1, a small air trimmer, is used for coarse calibration; I had planned to do fine calibrating electronically. Originally, C2 was a single polystyrene capacitor with the parallel combination of C1 and C2 capable of reaching 70 or 80 pF. The series-parallel modification of C2 shown in **fig. 7** resulted from a need for temperature compensation.

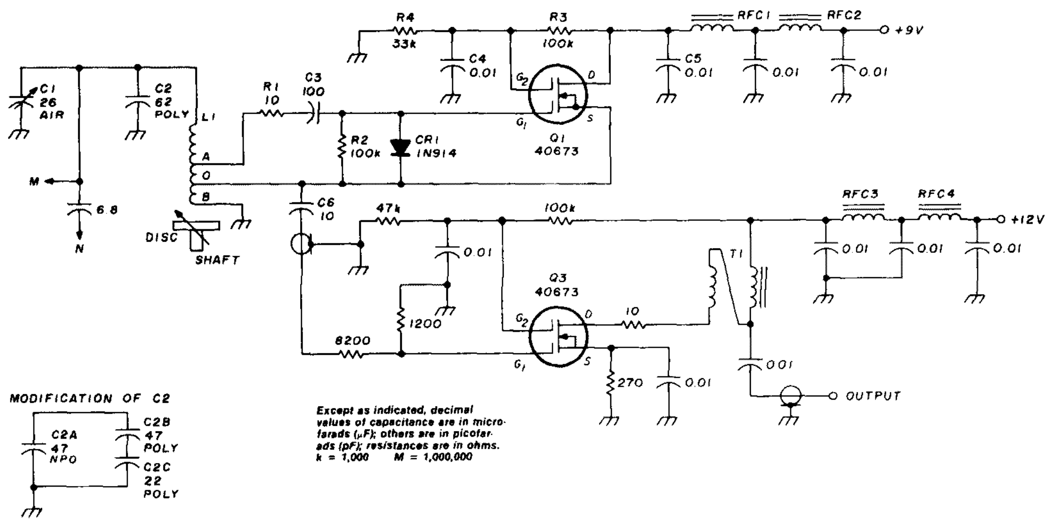


fig. 7. A disc-tuned, 30-MHz Hartley VFO and buffer for 15-meter use. L1, see text, T1 = 4:1 bifilar transformer, 6 turns of No. 26 enameled wire on Amidon FT 37-63 core. Points M and N, see text. Most of the labeled components appear in mechanical layout of *fig. 8*.

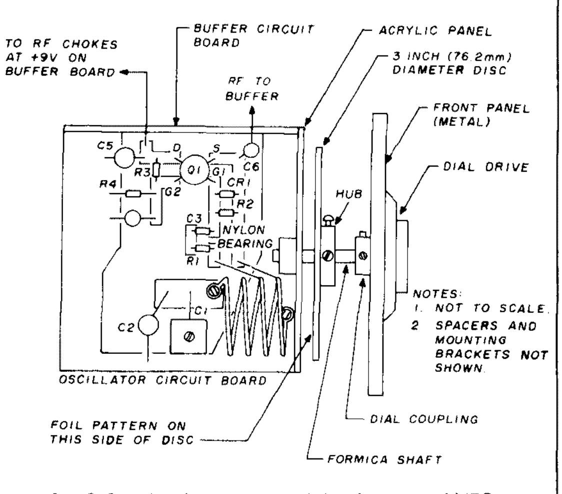


fig. 8. Mechanical outline of the disc-tuned VFO.

For the circuit layout, I used a single-side Fiberglas® epoxy-resin board. I prefer the etched circuit to free-style wiring because of the precision with which a layout can be planned and its mechanical stability. As for any objection concerning the relatively poor insulating properties of the board, I always etch the high-impedance conductors of an rf circuit with plenty of space between them and other conductors or ground. For stability during initial tests, and even in the shield box, a ground plate on spacers is a good idea.

In **fig. 7**, a tap-point M is where auxiliary capacitors can be switched in to make the oscillator run at 23 MHz for 20-meter operation. This entails a separate dial calibration, but for home construction, the method is much simpler than having a VFO with a single calibration, using an extra mixer, and switched-crystal oscillators to provide local-oscillator injection for each frequency band.

Tap-point N comes after a small coupling capacitor 6.8 pF. It is intended to be used for frequency slewing with varactors in the conventional manner. Possibilities include using a number of potentiometers, one for a shift of the VFO frequency in transmitting CW, one for vernier tuning in the receive position, and one for vernier frequency calibration adjustment.

Fig. 8 shows mechanical details of the VFO. Note that the buffer is mounted on a separate board at right angles to the main board and away from the coil. In its box, the outside dimensions of the unit measure 4 inches (101.6 mm) wide, 4 inches (101.6 mm) high, and 4-1/2 inches (114.4 mm) deep. The shielded box is actually intended to go inside the enclosure of a transceiver, instead of being tuned as shown in **fig. 8**, thus affording extra shielding.

The frequency calibration of the VFO is shown in **fig. 9** for two spacings of the disc from a reference point on the ground end of the coil. It should be easily possible to achieve band-spread tuning in excess of 1000 kHz, or as narrow as desired. Linearity is acceptable, but could be improved by tailoring the disc pattern.

The current to the VFO is 2.3 to 2.4 milliamperes as the disc is turned. The buffer draws 2 milliamperes.

temperature effects

In my early VFO work, I relied upon polystyrene rf capacitors and was wrongly inclined to blame a positive drift of frequency with temperature upon the air-core toroidal inductors I had developed.[1] Publication of careful work by other experimenters has since revealed that the polystyrene capacitors have a high negative temperature coefficient.[3,4] The type of capacitor recommended as an alternative is the NPO ceramic; these are available through the large mail-order wholesalers. At the time of this writing, I learned from W7ZOI that they could be obtained from at least one distributor: Mouser Electronics, 11433 Woodside Avenue, Santee, California 92071. Actually, I found that just inserting an NPO ceramic capacitor as C2 of **fig. 7** in my experimental unit was not satisfactory, since a negative coefficient of temperature with frequency resulted. I found it was not hard to compensate for frequency drift with temperature by mixing NPO ceramic and polystyrene capacitors, substituting the series and parallel combination shown in **fig. 7** as "Modification of C2."

The heat run shown in **fig. 10** was accomplished with the same plate warmer and 40-volt ac source as

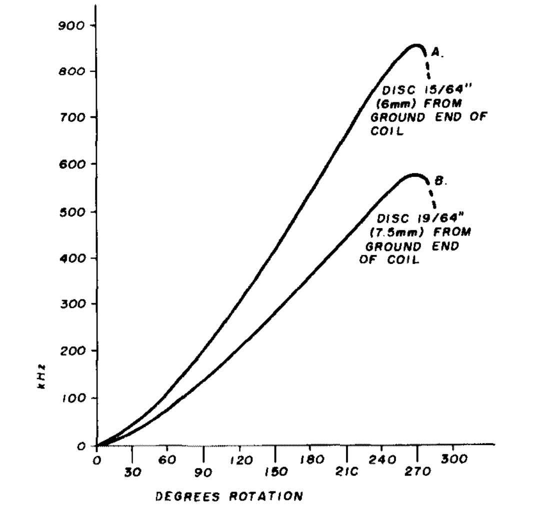


fig. 9. Incremental frequency calibration of 30-MHz disc-tuned VFO.

in reference 1. Results are not directly comparable, however, because of a different physical configuration and shield-box shape, with its obviously longer thermal time constant and greater heat losses than in the earlier case. Nevertheless, it is estimated that the temperature rise near the critical circuit elements approached 16 degrees C (29 degrees F).

The data for **fig. 10** were obtained by using 47 pF NPO for C2A; 47 pF polystyrene for C2B; and 22 pF polystyrene for C2C. The small dip in the curve for the first hour could indicate that the temperature transient reached C2A before C2B and C2C. It would have been interesting to make C2C an NPO capacitor.

In **fig. 10**, the frequency of the 30-MHz VFO has increased only 4300 Hz in fourteen hours on the plate warmer. The same VFO at constant room temperature for one hour did not appear to depart from its initial frequency by as much as 20 Hz.

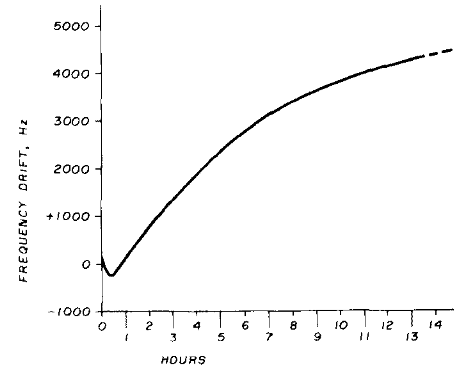


fig. 10. Frequency drift of the 30-MHz VFO during a heat test.

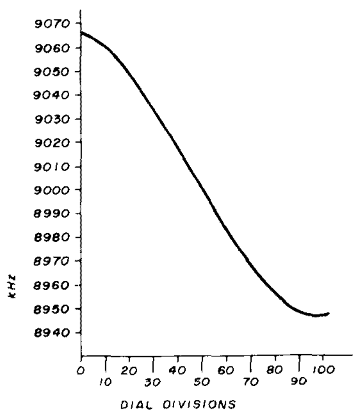


fig. 11. Corrected frequency calibration of copper-vane VFO. (See *ham radio* magazine, July, 1980, fig. 14, page 57.)

conclusions

The experiments reported here have shown that small, stable, low-current VFOs can be built at frequencies up to at least 30 MHz, using new versions of older inductance-tuning methods, eliminating the bulky and expensive variable capacitor, and now with cylinder- or disc-tuning, extending the useful tuning range beyond 270 degrees. Also, the wiping contacts found in most variable capacitors are eliminated.

The oscillators shown here could be made smaller. For the disc-tuned unit, one easy way to cut down on size would be for part of the disc to protrude through a slot in the shield box. Perhaps the ultimate compact oscillator would be a cylinder-tuned type using a Hartley circuit and perhaps a single closely-placed coil of the solenoid type near the circuit, operating at 30 MHz or higher.

One foreseeable problem with a 30-MHz VFO, especially the disc-tuned type, would be that of microphonics, especially if a loudspeaker were close by, or perhaps from vibration caused by keying a transmitter. One recalls the all-wave broadcast receivers of the 1930s, in which the tuning capacitors were mounted on rubber shock absorbers.

Appendix I demonstrates how the band-coverage capabilities of any inductance-tuned local oscillator go up with frequency. With stable, inductance-tuned VFOs of high enough frequency to be capable of a 500- or 1000-kHz tuning range, it should be possible to design very simple, general-coverage high-frequency receivers of fairly high quality.

appendix I

Here is a little mind sharpener for budding future engineers to demonstrate that mathematics is a useful and fascinating tool, and for over-the-hill engineers to convince themselves that they are still with it.

The oscillators I have described all tune through a band of frequencies by shifting the frequency of oscillation an incremental amount. The shift is accomplished by making incremental changes in the inductance of the tuned circuit. The question is: Assuming that the total inductance of the frequency-determining circuit is shifted by a certain small, proportionate amount, what is the relationship of the frequency change to the frequency itself?

First there is the fundamental equation relating frequency to inductance and capacitance:

$$f = 1/2\pi \sqrt{LC} \qquad (1)$$

where f is frequency in Hz, L is inductance in henrys, and C is capacitance in farads. Any other units are taken care of by the use of appropriate constants. Now, squaring **eq. (1)**:

$$f^2 = 1/4\pi^2 LC$$

If we keep the capacitance constant and change the inductance by a small amount, dL, the frequency changes by a small amount df, so we can write:

$$(f + df)^2 = 1/4\pi^2 C (L + dL)$$

Then expanding:

$$f^2 + 2f\,df + df^2 = 1/4\pi^2LC \times 1/(1 + dL/L)$$

Substituting f^2 from above for $1/4\pi^2LC$, and dropping the higher-order term df^2, which we can do for small increments:

$$f^2 + 2f\,df = f^2/(1 + dL/L) \tag{2}$$

Now for small dL/L:

$$1/(1 + dL/L) = 1 - dL/L$$

so, from **eq. (2)**:

$$f^2 + 2f\,df = f^2(1 - dL/L),$$
$$2f\,df = -f^2dL/L$$

and:

$$df = -f/2 \times dL/L \tag{3}$$

This equation says that, for small changes, if the change in the inductance is a fixed proportion of the inductance itself, dL/L constant, then the frequency change is proportional to the frequency being changed but in the opposite direction. This may be a fairly useful relationship when one is striving for the maximum achievable frequency shift. An analogous relationship is easily derived for capacitance tuning.

As the frequency-changing surface or object is brought closer to the inductive circuit, the inductance becomes smaller, but so does the Q, because the eddy-current paths in the metal are resistive as well as inductive. As noted before, when the Q is lower, oscillation eventually ceases, but before this happens, the oscillator may become a selective noise generator. Even when the oscillator appears to be acting normally, if the Q is too low there may be excessive noise modulation, making adjacent-channel signals audible with noise modulation.[5] This may be one kind of limit on the maximum achievable frequency shift. With cylinder tuning, the limit may sometimes be the fact that the cylinder's curvature does not allow close enough coupling to the coil.

The limitation on how big an incremental band can be covered by the methods outlined, and its relationship to frequency, may be more complex than indicated by **eq. (3)**. However, in general, the higher the local oscillator frequency the larger the band one can cover. If **eq. (3)** holds, then the 165-kHz available bandspread of the 9-MHz cylinder-tuned VFO of **fig. 4** could be scaled up to 3-1/3 × 165, or 550 kHz for a 30-MHz oscillator. Actually, I have converted the 40-meter direct conversion receiver that I described in the January, 1977, issue of *ham radio* to 15 meters, using a cylinder-tuned local oscillator. This oscillator could easily be made to function at 21 MHz with well over 1 MHz bandspread, because I coupled the entire oscillator inductor to the cylinder.

appendix II

Fig. 11 is a corrected version of **fig. 14** in reference 1. It should be noted by examining the new figure that the correct tuning range of the 9-MHz copper-vane VFO was about 120 kHz.

references

1. Richard Silberstein, WØYBF, "Variable-Inductance Variable-Frequency Oscillators," *ham radio*, July, 1980, pages 50-59.
2. R.C. Easton, K6EHV, "AFC Circuits for VFOs, *ham radio*, June, 1979, pages 19-23.
3. Wes Howard, W7ZOI, and John Lawson, K5IRK, "A Progressive Receiver," *QST*, November, 1981, pages 11-21.
4. Roy W. Lewallen, W7EL, "An Optimized QRP Transceiver," *QST*, August, 1980, pages 14-19.
5. B. Priestley, G3JGO, "Oscillator Noise and its Effect on Receiver Performance," *Radio Communications* (Radio Society of Great Britain), July, 1970, pages 450-457.

ham radio

low-power keyer
and interface

CMOS and MOSFETs in a keyer and switch that runs for two years off a small battery

A number of excellent, low-power, CMOS keyers have appeared in the literature in the past few years. Nearly all consume negligible power in the quiescent state, and would seem ideal for portable use or extended operation from a small battery, except for one problem. The interface between the keyer circuitry and the transmitter has always been implemented with either bipolar transistors or (reed) relays. These devices require more current to turn them on than the rest of the keyer circuitry.

I have been using a very simple, three-chip CMOS keyer (called HOPKEY MARK-IV) for over four years. This circuit has been duplicated by many hams who were intrigued by its simplicity and low-power consumption. As you can see from **fig. 1**, this keyer represents almost ultimate simplicity. It has no memory, iambic operation, or other frills. It just provides the basic requirements of self-completing dots, dashes, and spaces, and instantaneous operation on key closure (that is, gated oscillator). A side-tone oscillator was included, since, even with only three chips, enough gates were left over to make this a no-cost feature.

Before adding the required transmitter interface circuitry, the quiescent current drain of this keyer is unmeasurable on my equipment (less than 100 nanoamps). At 50 WPM, with the side-tone oscillator driving a high-impedance crystal earphone, it draws only 100 μA from a nine-volt transistor battery. This circuit keeps on working down to almost three volts, where it draws only 20 μA. Resistor values could be increased by an order of magnitude or more if even less current drain is desired. However, susceptibility to rf pick-up would increase. It is estimated that the battery would last its shelf life (two years or so) with such a light load. However, until recently, I have been plagued by an additional one to ten milliamps of current drain, just to drive the transmitter keying transistors. The recent development of power MOSFET transistors has changed that.

conventional keying circuits

Conventional bipolar transistors (and relays), when operated as switches, exhibit current gain. In other words, only a small amount of current is required for them to switch a larger current. For keying a transmitter with a positive (open circuit) potential at the key leads, a simple interface (**fig. 2A**) is normally used. The keyer power supply must furnish sufficient current through the base resistor to ensure that the

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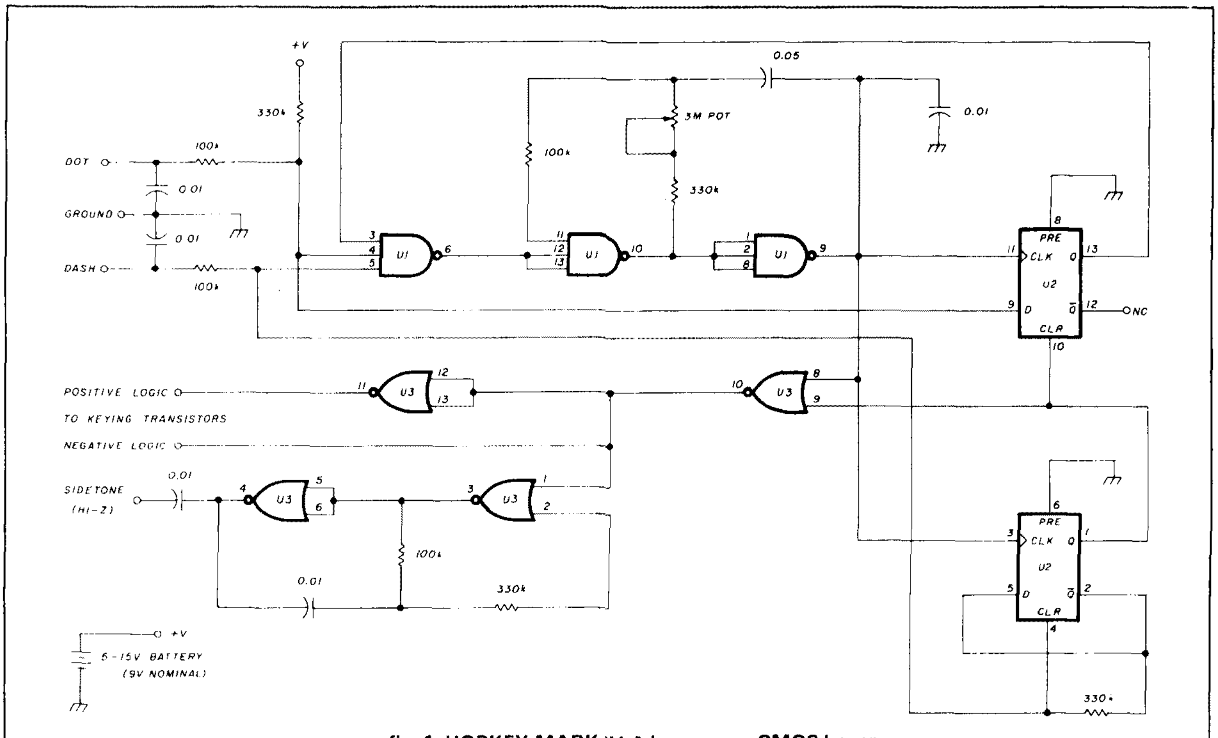

fig. 1. HOPKEY-MARK IV-A low-power CMOS keyer.

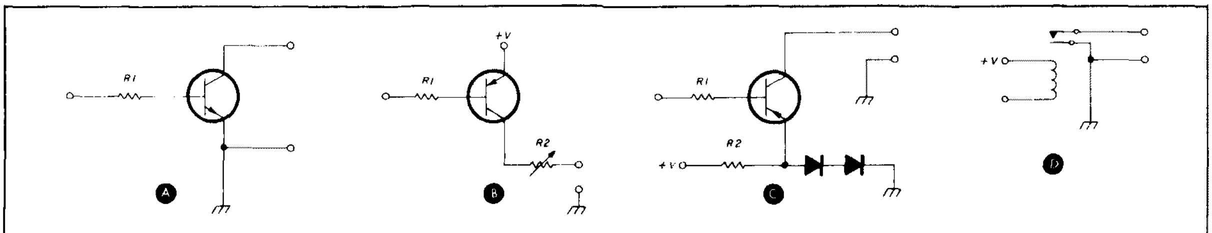

fig. 2. Conventional keying circuits. All draw considerable current from the keyer battery.

transistor turns on fully (saturates) with whatever collector current the transmitter presents in a key-down condition. This requires an additional 1 to 10 mA of current from the keyer battery.

For negative potential keying (grid-block), variations of the circuits shown in **figs. 2B** and **2C** are often used. These require a low-going input from the keyer, which sinks current from the base and turns on the transistor. Circuit 2B requires that R2 be adjusted for each particular transmitter in order to keep the key-down voltage near zero. The PNP transistors for grid-block keying normally have a high-voltage rating. Very often, 2N4888 or 2N5416 types are used since they will withstand −300 volts.

The simple relay interface of **fig. 2D** generally requires the highest amount of drive current, but offers the advantages of positive or negative keying, transmitter isolation, and relatively high-voltage operation.

All the above circuits draw current from the keyer power supply. While this can be as low as 1 mA in some cases, it varies with the type of transmitter to be keyed, and is generally many times greater than the current required to run the keyer logic. Battery life is severely shortened.

N-channel MOSFET keying circuits

Power MOSFET transistors, on the other hand, are almost perfect switches. A voltage rather than a current is required at the gate to turn them on. Their extremely high gate resistance draws only 1 to 100 nanoamps from the keyer. Recent power MOSFETs

table 1. N-channel power MOSFETs. Those marked with an asterisk contain internal gate-source diode.

manufacturer	type	BV_{DSS} (volts)	I_{DP} (amperes)	$R_{DS}(on)$ (ohms)	$V_{GS}(th)$ (volts)
Ferranti	ZVN0106A	60	0.38	4	2.4
	ZVN0120A	200	0.18	16	2.4
	ZVN0335B	350	1.04	3	3.5
	ZVN0365M	650	3.04	6	4.0
	ZVN0465M	650	6.4	2	4.0
	ZVN12A3M	30	16	0	2.4
	ZVN1304A	40	0.23	10	2.4
	ZVN1314A	140	0.16	20	3.0
International Rectifier	IRF123	60	5.0	0.4	4.0
	IRF220	200	4.0	0.8	4.0
	IRF322	400	11.0	2.5	4.0
	IRF420	500	2.0	3.0	4.0
	IRF450	500	10.0	0.4	4.0
	IRF513	60	2.5	0.8	4.0
	IRF710	400	1.2	3.6	4.0
	IRF820	500	1.5	3.0	4.0
Intersil	VN10KM*	60	0.30	4 (est)	2.5
	VN67AF*	60	0.60	8 (est)	1.2 (typ)
	IVN5000ANE	60	0.70	2.5	2.0
	IVN5200TNH	100	2.00	0.5	2.0
	IVN6000CNU	500	1.75	4.0	5.0
	IVN6200KNX	800	2.50	6.0	5.0
RCA	RCA9213A	100	1.0	2.50	?
	RCA9212B	150	5.0	0.30	?
	RCA9195B	150	10.0	0.15	?

(called VMOS, HEXFET, DMOS, SIPMOS, and TMOS by various manufacturers) are enhancement mode devices. Unlike their familiar linear MOSFET cousins used as rf amplifiers, these devices are normally off and turn on when the gate is biased in the direction of the drain potential. In this respect, they look more like bipolar devices, but consume (almost) no current. **Fig. 3** shows the sheer simplicity of a positive-potential MOSFET keying stage. All that is required is an N-channel MOSFET transistor. (The capacitor across the output simply keeps rf from getting back into the keyer.) However, a few precautions must be observed.

When not turned on, the transistor must be capable of withstanding the open-circuit potential presented by the transmitter. This is no great problem since transistors are available with BV_{DSS} breakdown ratings of 20 to 650 volts. When conducting, the transistor must handle the key-down transmitter current. Even small, inexpensive devices have I_{DP} practical current-handling capacities of many hundred milliamperes. Larger devices can switch over 4 kW. More serious concerns are the devices' on-state resistance and threshold voltage.

power MOSFET characteristics

A power MOSFET essentially acts like a variable resistor (or triode tube) and is characterized by a forward transconductance, g_{fs}. This is the ratio of drain-source current versus gate-source voltage. Fortunately, power MOSFETs have high g_{fs} (typically 0.2 to 4.0 mhos). This means that very large currents may be switched with low gate voltages. However, there is a limit to how low the on-state resistance, $R_{DS}(on)$, (from drain to source) can be made. This will determine the minimum voltage across the transmitter key leads in a key-down condition. In choosing a suitable device, one must pay attention to the $R_{DS}(on)$ rating at rated transmitter keying current I_K. The key-down voltage V_K will be:

$$V_K = I_K \bullet R_{DS}(on)$$

Another concern is that a minimum gate-source threshold voltage $V_{GS}(th)$ is required to operate a MOSFET. This is normally less than 3 volts and presents no major problem when the keyer is operated from 5 volts or greater. The positive-logic (high-going) output of a CMOS (or TTL) keyer is compatible and may simply be connected directly to the gate.

In some applications, although not here, two other factors become important. These are gate-source breakdown and drain-source breakdown due to voltage spikes. These factors are particularly important when using high-impedance drive circuits and/or switching inductive loads. Reference 3 provides a good discussion of the sources of these problems and effective cures. For resistive switching, simply driving with a relatively low-impedance such as CMOS at less than 20 volts is normally satisfactory. The rf bypass capacitor at the output also limits the rate of change of voltage which offers another measure of protection.

Keep in mind that power MOSFETs are high-impedance devices. Handling precautions similar to those when using CMOS apply. Store in anti-static containers, do not handle by leads, and use grounded soldering equipment.

Table 1 lists a number of N-channel power MOSFET devices suitable for positive keying applications (the references should be consulted for complete listings). It should be noted that some types (marked with an asterisk in **table 1**) have internal reverse diodes from gate to source. With such devices, the gate must never be allowed to go negative with respect to the source; otherwise the device can self-destruct. This situation cannot occur in the present application.

Power rating is generally of little concern in switching applications since we either have high voltage (off) or high current (on), but not both simultaneously.

Some particular devices I have used with great success are the miniature (TO-92) ZVN0106A and VN10KM for keying solid-state QRP transmitters and an ICOM-211. These devices cost less than a dollar. A very low key-down voltage of less than 20 mV is a particular benefit to some modern rigs (such as the IC-211), which balk at more than 0.3 volt across the key leads. The IVN67AF was used to key the emitter of a 10-watt solid-state amplifier. The IVN6000CNT was employed to cathode key an older rig using a pair of 6146s.

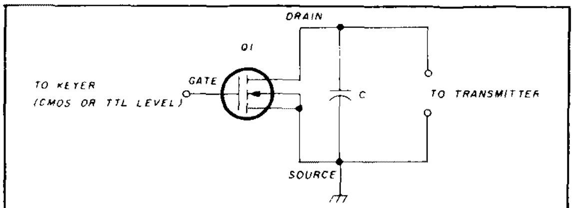

fig. 3. Positive-potential MOSFET keying stage. Capacitor C is 0.001-0.01 μF rf bypass; Q_1 is N-channel MOSFET.

P-channel MOSFET keying circuits

Keying a grid-block transmitter (negative potential) becomes a bit more involved, yet the same benefits may be achieved. Ideally, we would like to use a P-channel rather than an N-channel device. Some suitable P-channel devices are listed in **table 2**. To turn on a P-channel enhancement-mode, MOSFET requires a negative voltage at the gate. Although this would seem to make them incompatible with CMOS keyers, the circuit in **fig. 4** does the trick.

The input to this keying stage is the same positive-logic (high-going) signal from the keyer as was used above. It is inverted and applied to a voltage-converting circuit composed of a capacitor, resistor, and diode. When the keyer is at rest, the capacitor charges to almost the supply voltage (normally 5 to 9 volts). The diode provides a fast charging path. The resistor keeps the gate *near* ground once the capacitor is charged. When a dot or dash arrives, the inverter output goes low and the (negative) capacitor voltage appears across the gate-source. This turns on the MOSFET which keys the transmitter.

A small amount of charge is removed from the capacitor during key-down time. However, the RC time constant is made very large compared to the longest dash. This ensures that the transistor stays on during the entire dash. When a space occurs, the capacitor rapidly regains its previous charge and is ready for the next cycle. With the values shown, the transistor can be kept on for over 5 seconds. For long tune-up

table 2. P-channel power MOSFETs. Fewer types are available than with N-channel, yet they cover a wide range of operating characteristics.

manufacturer	type	BV_{DSS} (volts)	I_{DP} (amperes)	$R_{DS}(on)$ (ohms)	$V_{GS}(th)$ (volts)
Ferranti	ZVP0110A	−100	−0.20	16	−3.5
	ZVP0120A	−200	−0.13	40	−3.5
	ZVP0330B	−300	−0.68	8	−4.5
	ZVP0345B	−450	−0.68	8	−4.5
	ZVP0530A	−300	−0.06	200	−4.5
	ZVP12A3M	−30	−12	0.4	−3.5
International Rectifier	IRF9520	−100	−4.0	0.6	−4.0
	IRF9530	−100	−7.0	0.3	−4.0

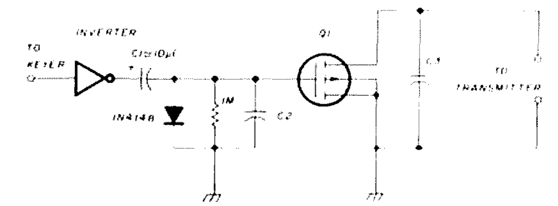

fig. 4. Negative-potential MOSFET keying stage. Q1 is P-channel MOSFET. C_1 should be low leakage. C_2 and C_3 are 0.001-0.01 μF rf bypass.

periods of greater duration, the key leads should be shorted directly rather than using the keying transistor to hold the transmitter on. The value of C_1 may also be increased to permit longer on time. Be sure that C_1 is a low-leakage type. A small capacitor (0.001-0.01 μF) from gate to ground may be required when operating with large amounts of rf (high SWR). This prevents the voltage-inverting circuit from acting as a positive peak-clipping rectifier which would keep the transistor turned on due to rectified rf. Very little energy is used in charging the capacitor since the high gate resistance drains off very little charge during key-down. Therefore, the power consumption of this circuit is only slightly greater than that of the one in **fig. 3**.

Use of a small (TO-92) type ZVP0120A permits keying up to −200 volts at 25 mA with less than 1 volt across the key leads in the on state. Only 20 mV appears across the key leads of my HW-100 with this circuit. ZVP0345B (TO-39 package) will key −450 volts at 100 mA with less than 0.8 volt across the key leads.

At last the ideal electronic switch seems to have arrived in the form of power MOSFET transistors. You can now run your keyer from a tiny battery for two years. Anyone for a solar-powered keyer? Or how about using rectified rf from the transmitter for power?

As you explore the benefits of these devices, you are sure to find many other applications for them. I have found that they even work quite well at rf as oscillators and power amplifiers. Why not get a few and experiment?

references

1. *VMOS Data Book*, 1981, Ferranti Limited. Available for $5.00 from Sales Offices and Franchised Distributors. Or, contact Ferranti Electric, Inc., Semiconductor Products Group, 87 Modular Avenue, Commack, New York 11725.

2. *Power MOS Handbook*, 1981, Intersil, Inc. 10710 N. Tantau Avenue, Cupertino, California 95014.

3. *HEXFET Data Book*, 1981, International Rectifier, 233 Kansas Street, El Segundo, California 90245. Available for $3.50.

ham radio

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SSB CONVERTER — CV-1982/TSC-26, 455 kHz. @ 50 Ohms input, Audio @ 600 Ohms output, Carrier, Subcarrier, Audio metered. Working Condition $80.00. N. E. Litsche, P. O. Box 191, Canandaigua, NY 14424.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. PO Box RY, Cardiff, CA 92007.

MX330 Motorola factory touch tone pad with mother board and daughter board, all interconnecting wiring. Nothing to cut or glue. Complete $100.00. N6GFE, 980 Wildcat Canyon Road, Berkeley, CA 94708 (415) 843-5253.

MANUALS for most ham gear made 1937/1970. Send $1.00 for 18 page "Manual List", postpaid. HI-MANUALS, Box R802, Council Bluffs, Iowa 51502.

SATELLITE TELEVISION — Howard/Coleman boards to build your own receiver. For more information write: Robert Coleman, Rt. 3, Box 58-AHR, Travelers Rest, SC 29690.

WANTED: Schematics-Rider, Sams or other early publications. Scaramella, P.O. Box 1, Woonsocket, RI 02895-0001.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

SATELLITE TELEVISION: Discount prices on all major TVRO items. Communications Consultants. (501) 452-3149.

RUBBER STAMPS: 3 lines $3.25 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

WANTED: New or used MS and coaxial connectors, synchros, tubes, components, military surplus equipment. Bill Williams, PO #7057, Norfolk, VA 23509.

VERY In-ter-est-ing! Next 5 issues $2. Ham Trader "Yellow Sheets", POB356, Wheaton, IL 60189.

CB TO 10 METER PROFESSIONALS: Your rig or buy ours — AM/FM/SSB/CW. Certified Communications, 4138 So. Ferris, Fremont, Michigan 49412; (616) 924-4561.

HAMS FOR CHRIST — Reach other Hams with a Gospel Tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

Coming Events

ACTIVITIES

"Places to go..."

INDIANA: The LaPorte Amateur Radio Club's Winter Hamfest, Sunday, February 27, LaPorte Civic Auditorium. Donation $2.50 at door; tables $1.00 each. Talk in on 52 simplex. For information/reservations SASE to: PO Box 30, LaPorte, IN 46350.

ILLINOIS: The Sterling/Rock Falls Amateur Radio Society's 23rd annual Hamfest, March 20, Sterling High School Fieldhouse, 1608 Fourth Avenue, Sterling. Tickets $2.00 advance; $2.50 door. Distributors, dealers and large flea market. Free parking and space for self-contained campers overnight. Doors open 7:30 AM. For tickets, tables and information: Sue Peters, 511 8th Avenue, Sterling, IL 61081 or call (815) 625-9262. Talk in W9MEP 146.25/85.

KENTUCKY: The annual Glasgow Swapfest, Saturday, February 26, 8 AM, Glasgow Flea Market Building, 2 miles south of Glasgow off highway 31E. Heated building, free parking, free coffee, large flea market. Admission $2.00, no extra charge for exhibitors. One free table per exhibitor, extra tables available at $3.00 each. Talk in on 146.34/94 or 147.63/03. For additional information: WA4JZO, 121 Adairland Ct., Glasgow, KY 42141.

MINNESOTA: The Midwinter Madness Amateur and Computer Fest sponsored by the Robbinsdale Amateur Radio Club, K0LTC, February 26, Sacred Heart Church School Auditorium, 4087 West Broadway, Robbinsdale. Public admission 8:30 AM. Seminars on antennas, towers, computer interlacing and a slide presentation on the voyage of the Viking ship, Hjemkomst. General admission $2.00 advance, $3.00 door. Commercial space $15.00 per table. Contact Bob Reid, N0BHC, 19725 Jackie Lane, Rogers, MN 55374. Flea market $3.00 per space. Contact Barry Blazevic, WB0FBN, 5437 Virginia Ave., N. New Hope, MN 55428.

MICHIGAN: The 21st annual Michigan Crossroads Hamfest sponsored by the Southern Michigan ARS and the Calhoun County Repeater Association, March 19, Marshall High School, Marshall. Doors open 7 AM for exhibitors and 8 AM for buyers and lookers. Tickets $1.50 advance, $2.00 door. Tables 50c per ft. For tables or tickets, SMARS, PO Box 934, Battle Creek, MI 49016 or call Chuck Williams (616) 964-3197.

MICHIGAN: The 13th annual Livonia Amateur Radio Club's Swap 'n Shop, Sunday, February 27, from 8 AM to 4 PM, Churchill High School, Livonia. Refreshments and free parking. Reserved 12 ft table space available. Talk in on 144.75/5.35 and 52 simplex. For information SASE to Neil Coffin, WA8GWL, c/o Livonia Amateur Radio Club, PO Box 2111, Livonia, MI 48151.

NEW JERSEY: The Old Bridge Radio Association's third annual electronic equipment auction, K of C Hall, Pine Street, off Route 18, Old Bridge. NEW LOCATION, plenty of seats and free parking. Doors open for registration and inspection at 9:00 AM; sale begins at 10:00. Admission $2.50. Club commission, successful sales only, 10% on first $100 of sale price, 5% on remainder. Refreshments available. Talk in on 72/.12, .34/.94 and .52. For information call Fred, WA2BJZ, (201) 257-8753.

NEW JERSEY: Shore Points ARC invites everyone to *Springfest '83, Saturday, March 12, 9 AM to 3 PM,* Atlantic County 4-H Center, Route 50, Egg Harbor City (near Atlantic City). Large heated building for buyers and sellers. Covered outside tailgating spaces. Admission $3.00 at gate; $2.50 advance. Sellers $5.00 per space (bring own table). XYL's and children free. Refreshments available. Talk in on 146.985 and .52. For info and reservations: SPARC, PO Box 142, Absecon, NJ 08201.

NEW JERSEY: The Split Rock Amateur Radio Association's sixth annual electronics auction, Friday, March 4, *V.F.W. Post #3401, State Route 53, Morris Plains.* Doors open at 7 PM for unloading and inspection. Auction starts at 8 PM. Admission $1.00. Club commission 10% on first $50 of each sale, above which a flat fee of $5 will be charged. Commissions cash only. Refreshments available. Talk in on WR2ADE, 146.385/146.985 or 146.52 direct. For information: SARA, PO Box 3, Whippany, NJ 07981.

NEW YORK: LIMARC, The Long Island Mobile Amateur Radio Club's indoor Hamfest, February 20, Electricians Hall, 41 Pinelawn Road, Melville, L.I., 9 AM to 5 PM. General admission $3.00. Advance sellers table only $10.00 to Hank Wener, WB2AIW, 53 Sherrard Street, East Hills, NY 11577 or 10 PM to Midnight (516) 484-4322. Refreshments available. For information: Sid Wolin, K2LJH (516) 379-2861 nights.

OHIO: The Mid Winter Hamfest/Auction and Flea Market, Sunday, February 13, Richland County Fairgrounds, Mansfield. Doors open 8 AM. Advance tickets $2.00; $3.00 at door. Advance tables $5.00; $6.00 at door. Talk in on 146.34/94. For information or tickets: Harry Frietchen, K8HF, 120 Homewood Road, Mansfield, Ohio 44906. (419) 529-2801 or (419) 524-1441.

OHIO: The Cuyahoga Falls ARC's 29th annual Electric Equipment Auction and Hamfest, Sunday, February 27, North High School, Akron, 8:30 AM to 4:00 PM. Tickets $2.50 advance; $3.00 door. Sellers bring own tables. Some available for $2.00. Talk in on 147.87/.27 or 146.04/.64. Details from CFARC, PO Box 6, Cuyahoga Falls, Ohio 44222 or phone K8JSL (216) 923-3830.

PENNSYLVANIA: The R.A.E. Eyeball QSO Party sponsored by the Radio Association of Erie, March 19, Perry Highway Hall south of I-90 on west side of Rt. 19. FCC exams. Send Form 610 to FCC Buffalo office by February 22. QSL drawing, bring your card. Reserved tables only, $3.00 per 8 ft. table. Admission $2.00. Talk in on 01-61-22-82. Refreshments available.

PENNSYLVANIA: The Lancaster Hamfest, Sunday, February 20, Guernsey Sales Pavilion, U.S. Route 30 East, Lancaster, 0800 to 1600. Dealer set up 0600 by reservation. Tables $10 in main display area, $6 in annex. General admission $3.00. Tailgating weather permitting. Talk in on 146.01/61, 147.615/015, 146.52. Send reservations to Hamfest Committee, RD #1, Box 56V, Blue Ball, PA 17519. Checks payable to SERCOM, Inc.

VIRGINIA: The 10th annual WINTERFEST '83 celebrating the 20th anniversary of the Vienna Wireless Society, February 27, 8 AM, Community Center, 120 Cherry Street, Vienna. CW contest, manufacturers and dealers, indoor flea market, outdoor Frostbite tailgating. Tables $5 and $10. Free parking. Tickets $4.00. Good food available. Talk in on 146.31/91 and 146.52 simplex. For information SASE to Winterfest '83, PO Box 418, Vienna, VA

22180 or call Jeff Wilkes, W4NFA (703) 281-4249 or on Virginia Sideband Net.

OPERATING EVENTS

"Things to do..."

FEBRUARY 12: YL-OM Contest, Saturday, February 12, 1800 UTC to Sunday, February 13, 1800 UTC and Saturday, February 26, 1800 UTC to Sunday, February 27, 1800 UTC. OMs call "CQ YL" — YLs call "CQ OM". All bands may be used. No cross band operation. EXCHANGE: Station worked, QSO number, RS(T), ARRL section or country. Phono and CW scored as separate contests. Submit separate logs. Score one point for each station worked. Multiply number of QSOs by total number of different ARRL sections and countries worked. Contestants running 150 watts or less on CW and 300 watts PEP or less on SSB at all times may multiply the results of (C) by 1.25 (low power multiplier). Logs must be signed by operator and postmarked by March 15, 1983. No logs will be returned. *For further information contact* WA6WZN.

MARCH 12, 13, 14: Idaho QSO Party sponsored by the Kootenai Amateur Radio Society, 0000Z through 2359Z. Exchange RS(T) and county for Idaho stations; state and country for all others. Idaho stations score one point for each QSO multiplied times Idaho counties, states, VE provinces and countries worked. Others score one point for each Idaho QSO multiplied times total Idaho counties worked. Frequencies: CW - 50 kHz up from lower band edge. Novices 25 kHz up from their lower band edge. SSB: 3.920, 7.260, 14.250, 14.325, 21.325, 21.380 and 28.550. No net frequencies. Awards will be issued to top scorer in each Idaho county, state, VE province and country. USA mailing deadline for all entries is April 16, 1983. DX countries and Canada deadline is May 1, 1982. Send to *Vladimir J. Kalina*, KN7K, South 1555 *Signal* Point Road, Post Falls, Idaho 83854.

FEBRUARY 5, 6: Vermont QSO Party sponsored by the Central Vermont Amateur Radio Club (W1BD). 2100Z February 5 to 0700Z, February 6. 1100Z to 2400Z, February 6. Send SASE for official log and score sheets. SASE for results. Send logs/facsimiles, name, class of license, address, NLT March 1, 1983 to: D. Nevin, KK1U, W. Hill, Northfield, VT 05663.

FEBRUARY 13: The Oregon *Tualatin Valley* Amateur Radio Club will operate a commemorative special event station celebrating the 124th birthday of the state of Oregon, the 33rd state to be admitted to the Union on Valentine's Day in 1859. KA7CPT will operate from 1700Z to 0300Z on or near 14.280, 21.360 and 28.510. An attractive certificate QSL will be awarded to Amateur contacts who qualify. Send 9 × 12 SASE or $1.00 to Marshall D. McKillip, 1175 NW 128th Street, Portland, Oregon 97229.

FEBRUARY 15, 16: International DX Contest sponsored by the America Radio Club. Contacts with a Club DX Group member must be made during 0400 UTC, February 15 to 2400 UTC, February 16. Suggested frequencies: All authorized frequencies 10, 15, 20 and 40 meters, phone and CW. For a special award send QSL and $2.00 U.S. or 3 IRCs to America *Radio* Club QSO Contest, PO Box 3576, Hialeah, FL 33013.

PROJECT OSCAR, Inc. is preparing a new set of orbital predictions for the period covering the calendar year 1983. The predictions will provide the UTC times and longitude for all south to north equatorial crossings of AMSAT OSCAR 8 (AO8) and the 4 Russian satellites carrying transponders (RS5, RS6, RS7 and RS8). Minimum donation of $10.00 for mailings to the US., Canada and Mexico ($12.00 overseas). Send your name and address along with a check or money order payable to Project OSCAR, Inc. The donation covers the cost of first class mailing within the U.S., Canada and Mexico and airmail printed matter to overseas destinations. Project OSCAR, Inc., POB 1136, Los Altos, CA 94022.

WOULD LIKE TO GET IN TOUCH with other hams who are involved in emergency services, paid or volunteer, particularly those in emergency medical services (EMS) or those who are EMTs or Paramedics or equivalent. Please contact Jeff Howell, EMT, WB9PFZ, PO Box 463, Madison, IN 47250.

FORMING A NATION-WIDE NETWORK of motorcycling Amateur Radio operators. Anyone interested please check in on 3967 kHz at 0300Z Thursday evenings. Everyone welcome. Please SASE for details. Gary McDuffie, AG0N, Route 1, Box 90-A, Bayard, Nebraska 69334.

WORKSHOP: Personal Microcomputer Interfacing and Scientific Instrumentation Automation. March 21-24, 1983. *$595.00. The workshop is hands-on with participants designing and testing concepts with the actual* hardware. For more information, call or write Dr. Linda Leffel, C.E.C., Virginia Tech, Blacksburg, Virginia 24061. (703) 961-4848.

ham radio TECHNIQUES

Bill Orr W6SAI

Welcome to the new 10-MHz band! For those contemplating operation on this new portion of the spectrum opened in late 1982 to Radio Amateurs, here's some information that may be of interest.

experimental tests run in 1982

During 1982, several U.S.A. Amateurs had the privilege of conducting tests on the 10-MHz band using experimental licenses. The license is granted under Part 5 of the FCC rules for the Experimental Radio Service. This was not an Amateur license, and communications between experimental stations and Amateur stations were not permitted. Licenses were granted by the Office of Science and Technology of the FCC on proof of necessity, with strict requirements concerning frequency, power output, and operating practices (**fig. 1**).

The purpose of the ensuing tests conducted over a period of a year was to determine the characteristics of the 10-MHz band, to test various antenna designs for the band, and to see if Amateur-style operation was feasible among the various commercial stations occupying the band. Power limit for the experimental stations at first was very low, but was gradually raised as I gained operating experience. Now that the band is open for general operation in the United States, the need for the experimental 10-MHz transmissions has ceased and plans are afoot for investigation of the future 18 and 24 MHz Amateur assignments.

A formal report on 10-MHz experiments will be filed with the FCC, but the investigations of KM2XDW (W6SAI) may be of general interest to the readers of this column.

10-MHz operating conditions

After conducting tests across the United States and receiving overseas reception reports from Europe, Africa, and Australia, aided by hundreds of hours of monitoring the band, I've come to the conclusion that 10 MHz resembles 7 MHz more than it does 14 MHz. Long-distance DX is workable on 10 MHz much as it is on forty. Some mouth-watering signals have been logged: FB8WG, VK9YC, ZS, VK, and ZL, plus stations in Malta, Greenland, the Philippines, Japan, Indonesia, and South America. Over forty-five countries were noted as the year progressed and Amateur activity increased.

Even so, there were long periods of time during daylight hours, particularly in summer, when no signals were heard, aside from the 40-kW RTTY signal of NAA (Cutler, Maine) on 10.130 MHz. Summer static levels were high (compared with 20 meters), and only during hours of darkness was the band open for long-distance communications.

During the winter months, on the other hand, the 10-MHz band opened to Europe, Africa, and South America as early as 2200Z in the afternoon (in California). Most signals were weak, as the DX stations seemed to be running 100 watts or less into makeshift antennas. On the other hand, VK9YC, using 100 watts into

FEDERAL COMMUNICATIONS COMMISSION
EXPERIMENTAL
RADIO STATION CONSTRUCTION PERMIT
AND
LICENSE
(AS MODIFIED)

EXPERIMENTAL (RESEARCH) (Nature of service) — K M 2 X D W (Call sign)

EXPERIMENTAL XR EX (Class of station) — 8829-ER-ML-82 (File number)

NAME WILLIAM I. ORR

Menlo Park (San Mateo) California - Lat. 37 26 20 N; Long. 122 10 42 W.
(Location of station)

(Location of authorized remote control point)

Subject to the provisions of the Communications Act of 1934, subsequent acts, and treaties, and all regulations heretofore or hereafter made by this Commission, and further subject to the conditions and requirements set forth in this license, the licensee hereof *is hereby authorized to use and operate the radio transmitting facilities hereinafter described for radio communication.*

Frequency	Emission Designator	Authorized Power (Watts)	Special Provisions
10.1-10.15 MHz	0.1A1/3A3A	100 (ERP)	
18.068-18.168 MHz	0.1A1/3A3A	100 (ERP)	
24.890-24.990 MHz	0.1A1/3A3A	100 (ERP)	

In lieu of frequency tolerance, the occupied bandwidth of the emission shall not extend beyond the band limits set forth above.

Operation: In accordance with Sec. 5.202(b) of the Commission's Rules.

Special Conditions:

(1) The provisions of Section 5.155 are hereby waived.

The above frequencies are assigned on a temporary basis only and are subject to change at any time without hearing.

This authorization is granted subject to the condition that no harmful interference is caused to any other station or service and may be cancelled at any time without hearing if, in the judgment of the Commission, such action should be necessary.

This license is issued on the licensee's representation that the statements contained in licensee's application are true and that the undertakings therein contained, so far as they are consistent herewith, will be carried out in good faith. The licensee shall, during the term of this license, render such service as will serve public interest, convenience, or necessity to the full extent of the privileges herein conferred.

This license shall not vest in the licensee any right to operate the station nor any right in the use of *the frequencies designated in the license* beyond the term hereof, nor in any other manner than authorized herein. Neither the license nor the right granted hereunder shall be assigned or otherwise transferred in violation of the Communications Act of 1934. This license is subject to the right of use or control by the Government of the United States conferred by Section 606 of the Communications Act of 1934.

This authorization effective July 21, 1982 and will expire 3:00 A.M. EST November 1, 1983

7-21-82

FEDERAL COMMUNICATIONS COMMISSION

fig. 1. The U.S. Experimental License. The license is granted upon demonstration of need. It is a temporary grant, subject to change or withdrawal at any time without hearing. The authorization is granted subject to the condition that no harmful interference be caused to any other station or service. The license for KM2XDW authorizes operation on the 10.1, 18.068, and 24.890 MHz bands for CW and SSB emission with an effective radiated power of 100 watts (later modified to 600 watts output for CW and 1200 watts PEP output for SSB).

an inverted-V at about 40 feet, was as regular as clockwork almost every afternoon on the West Coast via the long path. Sometimes he was accompanied by VK6 signals from Western Australia. And these DX catches were mixed in among plenty of signals coming short path from Europe!

In the morning hours, around sunrise, the 10-MHz band was wide open to the Orient. When the Japanese Amateurs were finally allowed on the band, several were heard running only 10 watts into a loaded 20-meter dipole. And a few signals from Indonesia and Australia banged in, too.

Just below the 10-MHz band several out-of-band broadcasting stations in Vietnam could be heard. These served as excellent propagation check points for the Asian opening.

10-MHz antennas

Most of the 10-MHz Amateur stations heard during 1982 used simple makeshift antennas — center-fed antennas, inverted-Vs, long wires, and the like. One Scandinavian Amateur had a large V-beam (intended for 40-meter operation) aimed at the United States, and his signal was an outstanding one from Europe. No doubt some DXer will come up with a 30-meter rotary beam one of these days!

The experimental license afforded me an interesting opportunity to check simple antennas, as there was no interference if the operating frequency was carefully picked. During good conditions, contact could be established from California to New York on 10.125 MHz at almost any hour of the day or night.

Each station had two antennas that could be quickly interchanged — KM2XDW in California had an inverted-V with the apex at about 50 feet and a quarter-wave ground plane whose base was about 12 feet above ground. The ground plane had eight radial wires. These specific antennas were chosen as representative of typical, inexpensive types that were well-suited for 10-MHz service. KM2XDU (W2LX) in New York had a dipole at about 45 feet and a similar ground plane at the same base elevation as that of KM2XDW.

Over the California/New York path the inverted-V and the dipole were invariably better than the ground planes by 3-6 dB. In addition, man-made noise was appreciably lower on the horizontal antennas than the verticals. KM2XDW ran listening tests on European signals and also on VK9YC (Cocos-Keeling Island), and in all instances the inverted-V provided a more readable signal than the ground plane. The conclusion I reluctantly reached was that a ground plane antenna is satisfactory, but a simple dipole or inverted-V whose center is a half-wavelength high, or more, is a better antenna.

Ground conductivity in the vicinity of the station appeared to enter the

picture. KM2XDU (W2LX) seemed to feel that his ground plane was on a par with the dipole as far as reception went. His ground conductivity was very good, with the water table just below the surface. At my station, where ground conductivity is poor, results obtained with my ground plane were not impressive. This points up the interesting idea that good ground conductivity may play a large part in cases where a DXer has had above-average results with a vertically polarized antenna.

point-to-point operation

The frequent schedules between KM2XDU, KK2XJM (Florida), and KM2XDW reminded me that Amateur communications is generally a random operation. It is usually possible to contact somebody somewhere, unless the band has dropped out. Point-to-point operation is entirely different. The stations are locked in a specific route and if that propagation path isn't open, no communication exists, as there isn't anyone else to talk to.

High power and beam antennas can make a questionable path worthwhile. Many times the 100-watt-plus dipole signal of KM2XDU would be running S-zero in California and the 40-kW-plus-beam signal of NAA in Cutler, Maine (not many miles from KM2XDU), would still be very clear at S7 to S9.

Monitoring other foreign Amateurs pointed up the fact that 100 watts and a dipole antenna were sufficient for plenty of good DX operation on 10 MHz, and KM2XDW in California received good reception reports from Europe when running that power level during his one-way transmissions.

SSB or CW on 10 MHz?

The experimental stations had the luxury of running SSB transmissions back and forth, and no problems were encountered. But the practicality of SSB could come into question when the 50-kHz-wide band becomes more populated. How much of the band can be allocated to SSB transmissions? During daylight hours, there's no reason why the whole band can't be opened to SSB as the DX opportunities are few. But at night, when long distance contacts (and long distance QRM) abound, SSB transmission doesn't seem very practical. Perhaps a temporary U.S.A. authorization of SSB transmission from, say, 1400Z to 2200Z may be the answer. Amateurs have never had a general time restriction on a band, and perhaps this is the ideal chance to try one out. In any event, it might be a good idea to avoid contest-style operation on this band, at least until Amateurs get a feel for the operating conditions in this narrow sliver of the spectrum.

the Swiss cheese effect

An interesting "operating hazard" became apparent shortly after the 10-MHz point-to-point tests were started; it was immediately called the Swiss cheese effect. It had been noted before on other bands, but not to the degree apparent at 10 MHz. The effect was simple — during a contact signals would rapidly drop out for a period of seconds or minutes, then build up to normal strength again. The Swiss cheese effect was different from the type of fading normally encountered; it seemed almost as if a hole had opened in the ionosphere and the signals had somehow fallen into it. Sometimes the ionospheric hole lasted for only seconds, at other times it lasted up to three or four minutes.

It has been suggested that the ionospheric hole could be avoided by moving transmitter frequency a few tens of kilohertz, insofar as the hole may be frequency sensitive. Tests are underway to determine if this is so.

If these ionospheric holes exist, they might explain the mysterious and frustrating situation where a DXer seemingly cannot contact a faraway station, when other Amateurs in his vicinity and with comparable equipment seem to work the station with ease.

putting the Collins KWM-2 and S-Line on 10 MHz

Some of the newer pieces of equipment are ready to go on 10 MHz now, or can be put in the transmit mode by a simple modification. Older equipment, however, may take extensive modification to reach the new band.

The Collins KWM-2 and S-Line, happily, fall between these extremes and can be made operative with only a little effort by the owners. The following data applies to the KWM-2 specifically and to the S-Line generally.

For either model, new conversion crystals are required; one for the KWM-2 and two for the S-Line.* The 20-meter range is used for 10 MHz, and I placed the new crystal in the old WWV position (14.8 to 15.0 MHz). This left the 20-meter ham band intact.

It is a good idea to put a small label marked 10.0-10.2 MHz on the bandswitch so you won't get mixed up changing bands. Once the crystal is installed, the exciter tuning control is adjusted to approximately 3.1 and then peaked for maximum background noise. PA tuning is approximately 3.3.

The transmitter is now ready to be tested. Since the output amplifier had been adjusted for 14 MHz operation, it requires some additional tweaking to permit proper loading at 10 MHz. As is, the amplifier stage may be overcoupled to the antenna at 10 MHz and additional output capacitance in the amplifier pi-network is required for efficient operation (**fig. 2**). The capacitors in question are C-155 and C-152 (see instruction manual). These are mica compression types located on the chassis near the two control relays at the rear of the deck. They may be adjusted either from the top or from under the equipment.

Capacitor C-155 is permanently in the circuit and is normally adjusted on the 10-meter range so that the main loading control reads 50 ohms when a 50-ohm load is attached to the antenna receptacle. Capacitor C-152 does the same job on the 20-meter range and is switched into the circuit by

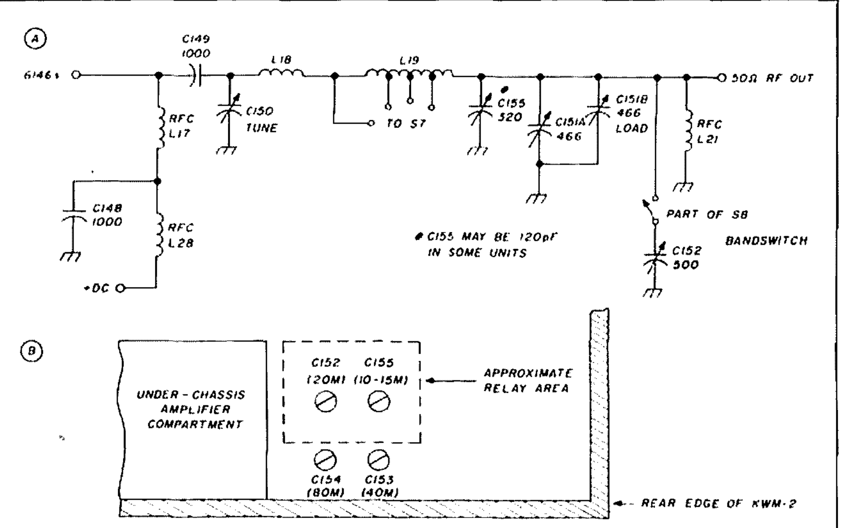


fig. 2. (A) Schematic of the KWM-2 transmitter amplifier stage. (The 32S-3 is similar but component nomenclature may be different, as well as parts placement.) The capacitors to be readjusted for 10-MHz operation are C-152 and C-155, mica compression units mounted on the chassis deck. (B) Under-chassis view of KWM-2 showing placement of compression capacitors near amplifier compartment and rear edge of chassis. Capacitors may be adjusted from above or below the chassis.

segment S-8 of the bandswitch. Note that in some early equipment C-155 may have a maximum value of 120 pF, and in later equipment C-155 may have a capacitance at 320 pF. If yours has the lower value capacitance, it will be very difficult to achieve proper loading on 10 MHz.

In any event, the technique is to increase the capacitance of C-152 to maximum value to achieve proper loading when the equipment is operated on the 30-meter band. If overcoupling still exists, then capacitor C-155 is adjusted to maximum value. (For normal operation on 20, 15, and 10 meters, these capacitors must be returned to their original settings.)

To eliminate the necessity of making these adjustments, an auxiliary loading capacitor may be connected directly across the antenna receptacle of either the S-Line or the KWM-2. A 350-pF broadcast-type capacitor will usually do the job. If additional loading capacitance is needed, a 200-pF, 1-kV mica capacitor can be paralleled with the variable capacitor.

The final stage is to realign the small variable padding capacitors in the receiver rf section for maximum gain at 10 MHz. This may be done by ear. If the 14-MHz setting of these capacitors is marked on the chassis with a pencil, it will be but a matter of seconds to realign the receiver to 14 MHz.

Other equipment, such as Drake, can be placed on the new band with the addition of a conversion crystal. However, it may still be necessary to realign the receiver section for maximum gain at 10 MHz and to determine if the pi-network of the transmitter will load into a 50-ohm system before you go on the new band with your first CQ call.

I'd be pleased to hear from our readers about how they get their equipment working at 10 MHz. If there are any interesting problems, I'll be happy to print them in this column for the benefit of all.

ham radio

*The crystal frequency for the 10.0 to 10.2 MHz range is 13,155.00 kHz (Collins part number 290-9042-000). The crystal can be obtained directly from Rockwell Collins Service Center, 920 Shaver Road, N.E., Cedar Rapids, Iowa 52498, attention Jim Maccani, W0HUP; telephone 319-395-5391.

the Bobtail curtain and inverted ground plane part one

History and useful information by the originator of this popular DX antenna

Woody Smith, W6BCX, the originator of the Bobtail Curtain, provides humorous and informative anecdotes on this popular DX antenna, using a Q & A format. Some of our older readers will recognize him as the previous editor of *Radio* (predecessor to *CQ* magazine). This article is well worth reading carefully. **Editor.**

I was flagged down recently at the monthly TRW (Los Angeles) Swap Meet by an old timer I hadn't seen for twenty-five or thirty years.

"Hey, Woody, I'm sure glad you are wearing jumbo call letters. As I recall you used to be pretty sharp on antennas. The wife and I just retired to a place in the country with enough room for me to put up some decent antennas for a change, and I sure need some help.

"Over the last several years I've been reading lots of good things about a 40 and 75 or 80-meter array called the Bobtail Curtain that's supposed to do a real job on DX, and I'm thinking of putting one up for 75 meters. Do you know enough about the Bobtail to answer a couple of questions I haven't found answers to?"

"Well," I replied as I looked away and scraped a circle with my big toe in a futile attempt to feign modesty, "if I can't answer them authoritatively I deserve to be embarrassed. I wrote the original article on the Bobtail, back in 1948."

"Nineteen hundred *what* did you say?"

"It appeared in the April, 1948 issue of *CQ* under my name, with the title 'Bet My Money on a Bobtail Beam,'" I added. Then, seeing as how he was duly impressed with my credentials, I proceeded to answer his questions, all of which I had been asked before at one time or other.

Because certain questions have kept recurring over the years, a recap of those particular questions along with brief answers would seem to be in order. Also included are historical data on the evolution of the Bobtail from the inverted ground plane (IGP). The IGP has not received the recognition and popularity it deserves as a highly effective 40 and 75-meter omnidirectional antenna for long-haul DX. Then, for the benefit of those who always like to know all about the why, some additional details and information will appear in Part II of this article.

basic Bobtail Q & A

Q. My 40-meter Bobtail does an amazing job on DX compared to my old antenna, but I don't have room for a 75-meter Bobtail. What if I put up only half a Bobtail on 75, with two tails instead of three? How should I feed it?

By Woodrow Smith, W6BCX, 2117 Elden Avenue, Apt. 20, Costa Mesa, California 92627

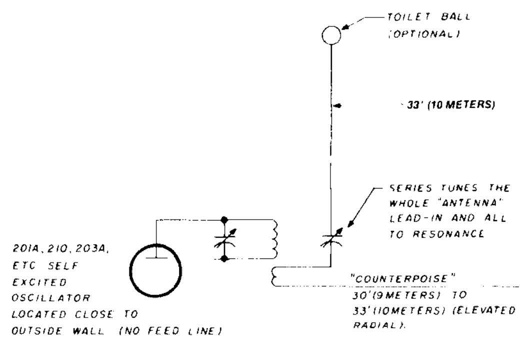


fig. 1. This simple 30 up and 30 out was responsible for lots of real DX on 40 back in the good old days. The copper toilet ball often was employed by the superstitious as a DX talisman (and by the author just for insurance). In 1928 the single-wing radial, then called a counterpoise, frequently was engineered to be about clothesline high (no driers in those days).

A. The three-element version is the elegant version, with better suppression of end-fire high-angle lobes from the horizontal section as compared to a two-element version without end radials. If end radials (extending out beyond the vertical elements) were employed on a two tailer, the horizontal space taken up would be the same as for the standard three-element Bobtail. A two-tailed version, by the way, actually is two-thirds of a three element, not half.

For gain, the two-tailed version without end radials (nowadays sometimes referred to as the half square, per K3BC) is nearly as good as the three tailer if properly fed. I'm partial to feeding the bottom of either leg via a resonant tank. Refer to the answer to the third question regarding coax feed.

Q. I'm going to have trouble getting poles up high enough on 75 meters. Can I cut off 15 or 20 feet from the tails of a 75-meter Bobtail by inserting loading coils in each tail near the bottom? If so, how far up should they be placed?

A. Yes, go ahead. On 75 I would place the coils up about 5 or 6 feet from the bottom. Don't shorten the poles and the tails any more than you have to, or the business part (top) of the vertical radiators won't be able to "see out" as well. Construction of suitable loading coils will resemble good quality trap coils. Any loss in performance other than a slight reduction in bandwidth will be a result of the lower antenna height. There will be very little loss in gain when using coils if the Q is reasonably good.

On 40 meters I see no excuse for loading coils. I would use poles at least 40 feet high so the current loops are well up off the ground, With poles this high on 40, no loading coils are required. If nearby buildings are more than one story, still higher is better yet. Keep in mind that the tops of the vertical elements always like it better when they can see out.

Q. Why can't I just feed one of the current loops of a two-tail Bobtail with coax? How well will it match 50 ohms?

A. You can feed it that way, and it will work, and the VSWR will be tolerable. The coax should be brought down at a 45-degree angle *toward the center*, not to the side or outside, until at ground level. Then take it where you want. There is no way to dress the coax that will avoid completely all undesirable coupling to the far side of the antenna, and this will result in some antenna effect on the coax. Fortunately, it will not be bad enough to cause serious problems. Unfortunately, coax will not work satisfactorily at half or twice frequency.

Q. When three vertical elements are used with bottom feed of the center element, how does the current compare in the three elements? Is it the same in all three, or twice as high in the center element? Or something in-between? I've heard arguments about this.

A. Intuitively one might conclude that the *current* distribution is 1-2-1 (binomial). But I learned long ago to be wary of deductions that are immediately obvious. What if the complex mutual impedances existing between the various elements should produce a significant effect upon the current distribution? These impedances and the net effect are quite difficult to calculate. The original article stated simply that the current is considerably greater in the center element. Measurements taken subsequently with the aid of a spotting scope confirmed that the distribution in a typical installation *approaches* 1-2-1.

Q. In your *CQ* article and in the description of the Bobtail in your book *The Antenna Manual* you show inductive (link) coupling between the feedline and the parallel resonant matching tank that voltage feeds the driven element. Can't I just use a variable tap instead, or maybe a tapped L-network? It would be easier to adjust than a link.

A. Inductive coupling was chosen primarily to cut down on possible receiver front-end overload and cross modulation. A 40-meter three-element Bobtail looks like a big omniverous Marconi to 160-meter and broadcast-band signals. If you don't have any 160-meter friends nearby or any high power a-m broadcast stations within a few miles, you should have no trouble using a tank or L-network with a variable tap on the coil (in lieu of the inductive link). You can always add a 50-ohm highpass filter designed for about

2500 kHz cutoff if you do happen to come up with a cross-modulation problem.

Q. In your description of the Bobtail Beam in the *Antenna Manual*, but not in your *CQ* article, you mention the use of a small ground screen under the bottom end of the driven element. How important is this? What are the benefits?

A. Such a screen makes a highly effective rf ground, much better than something buried in or driven into the soil, for a ground-independent antenna (meaning one which has little current flowing to ground or ground substitute at the feedpoint). The Bobtail falls in this category. Resonant radials above ground get in the way, are not required for efficient operation of a Bobtail, and may actually upset the pattern under some conditions.

An earth ground is useful primarily for lightning protection, and even if one is employed near the feedpoint for this purpose, a small ground screen in addition is recommended. Grounding considerations are covered in more detail in connection with further discussion of feed methods.

evolution of the Bobtail

The Bobtail may be considered as a broadside array of co-phased quarter-wave radiating elements configured as inverted ground planes. Let's start this Bobtail discussion with a review of the inverted ground plane before progressing to an array using them.

If you have trouble accepting a ground plane with only one radial, don't. Maybe the definition of ground plane has to be stretched a bit, but in the late 1920s (with some still in use in the early 1930s) there was a widely used 40-meter DX antenna often referred to as the 30-30 (**fig. 1**) which could be considered a ground plane flying on just one wing. It used a vertical quarter-wave radiator in conjunction with a neck-high quarter wave horizontal counterpoise which was nothing more than a single above-ground radial.

When the hams moved from 160 meters to 80 and then to 40, the easiest thing to get going in a hurry was a scaled-down antenna-counterpoise arrangement used on the lower bands. Usually the 30-foot radiator and the 30-foot counterpoise were brought in directly to the rig, placed by a window to keep the inside leads short. Feedline? Who needs a feedline? The overall length, with a sum total of about 60 feet outside, was just about right for series tuning to resonance by means of a variable capacitor, more often known in those days as a variable condenser.

Sometimes a copper toilet ball was placed atop the vertical radiator as a combination DX good luck charm and top-loading capacitance that substituted for the multi-wire flat top on a 160-meter Marconi. One big gun DXer claimed it put some kind of DX English on the radiated wave, while the small-caliber crowd always looked to see if he had tongue in cheek. Yes, I used a copper toilet ball. Just in case. No use taking any chances. Besides, nobody had proved yet that the ball did not do any good.

Don't ever pooh-pooh this venerable antenna, because its record of DX worked on 40 speaks for itself. Back in the late '20s a local ham friend worked (QSL confirmed) what was then Madagascar, now Malagasy Republic, on 40-meter CW a half hour before local sunset, running about 50-watts input. Yes, he did it with his trusty 30-30, complete with toilet ball. The rig used a 210 7-1/2 watt triode in a self-excited oscillator, and except for tube type, was typical of perhaps half the CW rigs on the air. Not too shabby from California, even if conditions did happen to be especially good at the moment. From a decent location and with good conditions such results then were commonplace enough with a 30-30 to be considered only slightly amazing.

Actually, the old 30-30 corresponds to a modern trap vertical that uses about 30 feet of effective vertical radiator on 40 meters working against an above-ground resonant radial. The toilet ball, when used, did add to the effective height, but without a loading coil probably not very much.

the center-tapped Windom

While the 30 up and 30 out was popular as a simple yet effective 40-meter DX antenna, the traffic and rag-chewing crowd on 40 had their very own favorite for short- and medium-haul work. This was the single-wire-fed Hertz, oriented horizontally at 30 to 40 feet. Its performance out to several hundred miles was such that its popularity and reputation were well deserved. And it was the ultimate in simplicity.

The antenna first got media attention in an article by Williams, 9BXQ (no W prefix back then), appearing in the July, 1925, issue of *QST*. This was followed by several others over the next few years.

As the name implies, this dipole antenna was fed by a single wire attached to a super-magic point on the dipole between 1/7 and 1/6 of the antenna length from the center. The exact point for minimum VSWR varied with feeder and antenna wire sizes and with surrounding objects, particularly ground.

This does minimize standing waves on the feeder, often bringing the VSWR very close to 1.0 if the dipole length also is correct. But contrary to a misconception widely held at the time (and still somewhat prevalent), unity VSWR does not eliminate radiation from (and pick-up by) the single-wire feeder.

Reduce radiation and pick up? Yes, some. Eliminate it? No. We have simply converted the line to a

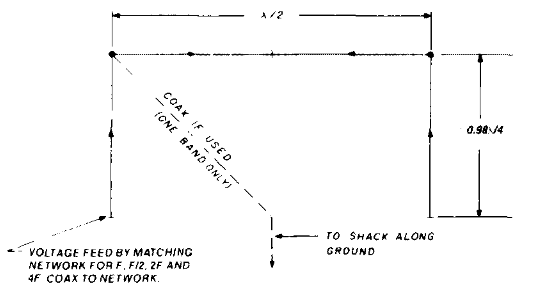


fig. 2. The old 30-30 works even better upside down. When inverted, two can be tied together as shown and voltage fed at one end to make an effective broadside curtain for DX. This elementary Bobtail, or half square did not stir up much interest back in 1948 because it looked too simple, but lately has received a lot of favorable attention.

traveling wave radiator (antenna). Minimizing the VSWR alters the pattern of radiation from the feedline somewhat, and reduces but by no means eliminates the net feedline radiation and pickup.

By 1929, enough conflicting information was floating around on the proper method of arriving at the magic tapping point for the feeder that Loren Windom, W8GZ, was prompted to write what has become a classic article on the subject. The article appeared in the September, 1929, issue of *QST*, and made it unnecessary to fret or argue over the subject any further.

Remember the Yagi-Uda situation, where the English-speaking Mr. Yagi (later Dr. Yagi) made it very clear in his classic 1928 IRE Proceedings paper that he was merely reporting on the work of Professor Uda, who had developed a clever new parasitic array a couple of years before? Well, the same thing happened with the single-wire-fed Hertz. Much of the early work was done at Ohio State University, and W8GZ gave them full credit. W8GZ made it very clear that he was acting solely as a reporter and was claiming no credit for collaborating on the actual development.

Nevertheless, over the years the single-wire-fed Hertz became better known as the Windom. In fact, in Great Britain it was generally referred to as the Windom almost from the day the article by W8GZ first appeared. Dr. Hideji Yagi, meet our Mr. Loren Windom, another reporter on antenna developments. He, too, unwillingly became world famous for an antenna he did not develop or invent.

Back when horizontal Windoms were common, an acquaintance of mine with one at 40 feet kept insisting that he could raise DX easier if he changed the match by sliding the tap a bit toward the center. He wondered if there were some easy way to figure out where the optimum DX tap should be attached without moving it a few inches at a time and comparing results (not too practical).

At first he thought I was kidding when, after getting suspicious as to what actually was going on, I suggested he move the tap to the exact center and see what happened. How about dropping the feeder straight down for about 33 feet, then cut it there and voltage-feed the bottom end with the Zepp feeders he had saved when he converted his Zepp to the Windom?

About a week later he called me breathlessly to announce that the new antenna was working so well that over the weekend he worked some new countries. He would have phoned me sooner except that he was too busy working DX, he explained.

upside down is better

On-the-air tests showed that this inverted configuration of what today would qualify as a two-radial ground plane consistently outperformed typical 30-30 installations on long-haul DX. Subjectively the improvement appeared to be at least a full S unit (then called an R unit).

Tests run more recently confirm that there is only one way to get a regular ground plane to perform as well as an inverted one. That is to get the whole ground plane up in the air where it is well removed from ground and pretty much in the clear. But on 40 and particularly on 75/80 meters this seldom is feasible.

Pat Hawker, G3VA, editor of the RSGB (Great Britain) book *Amateur Radio Techniques*, long ago recognized the advantages of turning a ground plane upside down at high frequency. For years Pat has been hawking (excuse me, *extolling*) the merits of the inverted ground plane for DX in his book.

the Bobtail takes shape

When it came time for me to get something back on the air after WW2, I recalled the results obtained from an inverted ground plane on 40-meter DX, and got to wondering: Is there something I could squeeze on my lot that would do a better job on 40-meter DX than an inverted ground plane? How about two of them in phase (**fig. 2**), oriented so the bidirectional pattern would cover the most important geography? How about using only one radial for each vertical element and bringing the radial ends together so that only two poles would be required? The half-wave spacing would be just right for broadside (in phase) operation of the vertical elements. And the voltage and phase at the tips of the two upstairs radials would be the same and therefore could be joined.

The antenna now would resemble nothing more than a bent fullwave antenna; so it should be possible to get away with feeding only one end (either end). The radiation from the two halves of the horizontal section should cancel well enough that the spurious end-fire lobes from the horizontal section don't represent much wasted power. Receiving, these minor lobes are going to pick up off-axis QRM, but it shouldn't be too high a price to pay for such simplicity.

With the project barely past the bill of materials stage, came an unsurmountable obstacle: I would be moving. There was nothing to do but abandon the project. The trouble was that having gotten all steamed up about the new brainchild, I just couldn't stand not having somebody, *anybody*, put one up to confirm my expectations. So I approached some of the local DX, golf, fishing, and self-styled world-class antenna experts and tried to interest at least one of them in putting up a 40-meter job.

Sad to relate, the very simplicity of the antenna turned out to be my undoing. No takers, even when I offered to help put one up. Their reaction was unanimous. They patiently pointed out to me that, as any fool could plainly see, no antenna that simple could possibly be much good, especially when it is upside down with the counterpoise on top. Obviously, if anything as simple as a bent piece of wire could be all that wonderful on DX, everybody would be using one.

How about enticing them with a more elegant version I had been thinking about. It would perform only slightly better and would require 50 percent more room, but would appear to be more sophisticated, more complicated, and more elegant looking. It definitely would not look like a bent piece of wire. How about adding a vertical element and feeding the bottom of the center one? It would produce only slightly more gain, but a cleaner pattern. More important at the moment, it would certainly be more impressive-looking when sketched.

Fortunately it did turn out to be easier to sell. I quickly got a willing customer who had room for a three-element 40-meter job. Thus, the Bobtail was born (see **fig. 3**).

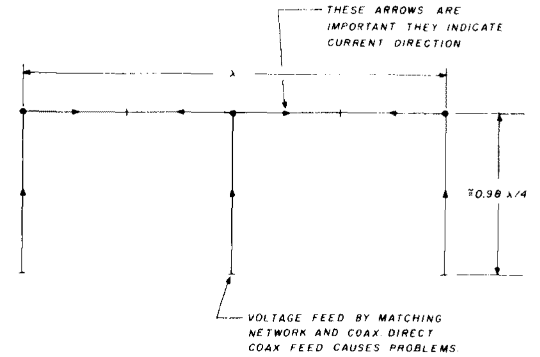


fig. 3. The classic three-element or elegant Bobtail provides better cancellation in the horizontal portion, resulting in less short-haul QRM when trying to receive weak DX. But it takes up more room (usually too much for 75 and 80 meters).

When he reported back to me on its DX performance, he kept using the words phenomenal, fantastic, etc., ". . . especially beyond 2500 miles when compared to my old antenna."

As a result of his plugging it over the air, I started receiving requests for information. To cut down on this I decided to write an article describing the antenna. When I contacted the editor of *CQ* about a Bobtail article, I recounted my lack of success in stirring up interest in a simple, two-element version. We decided not to include the two-tailer, but possibly make it the subject of a follow-up article.

The Bobtail with its three elements looked intriguing enough in the published article to inspire some readers with room to put one up to action. Then among some fan mail and requests for more information appeared a couple of surprises. Two correspondents advised me independently a few days apart that to get a Bobtail to fit their lots they had gone ahead on their own and made it more compact. Both did it by using two instead of three elements, and feeding one end of what was left. Both correspondents were quick to add that their simplified versions worked just great, gave fantastic results, etc., etc. "Just thought you might like to know."

I wrote them indicating I was glad to hear that their chopped Bobtails were doing such a good job, and congratulated them on their ingenuity. Somehow I felt it would appear pretentious of me to write an article on my truncated Bobtail, so never did.

Thanks to Ben Vester, K3BC, for seeing that it finally got some favorable publicity ("The Half-Square Antenna," *QST*, March, 1974). And speaking of the Half Square, Ben's designation certainly is tidier and more descriptive than something like Two-Element Chopped Bobtail Curtain.

Part II will include, among other things, quantitative information on the gain of the Bobtail and Half Square (both free-space theoretical and real world *practical* DX-signal gain), multi-band operation and performance, more information on feed methods, construction considerations, and some dimensions for 10-MHz Bobtails.

ham radio

NEW products

learning Morse

When we learned the Morse code, state-of-the-art was memory, and pencil and paper exercises: A- didah; B- dadididit. Over the past few years, records and tapes have come into vogue as learning aids. Now there are microcomputers which do the same. The Morsetalker MMS-1 from Spectrum International is one of these.

This interesting microprocessor-based device is a self-contained random Morse generator that incorporates a speech synthesis system. The MMS-1 has a speed range of 2 to 20 WPM in 2-WPM increments, and character group lengths of one, five, and fifty letters, before talkback. The unit is designed to work at six different learning levels: letters A-F; A-M; A-U; and A-Z; numbers only 0-9; and all letters and numbers combined.

The MMS-1 is designed to use the current teaching philosophy of sending at high speed with long spacing between letters. A crystal oscillator is used as a reference to ensure that all characters are sent and spaced accurately. For the more advanced Amateur, a high speed option is available to increase the speed range of the MMS-1 to 12-48 WPM in 4-WPM increments.

Using the MMS-1 is very easy. Push buttons are used to select character range, group length and speed, with LEDs indicating group length and speed. Once you have made your selection, push the go-stop button and you're all set to start.

There are a few minor drawbacks to the MMS-1. You cannot alter pitch or volume, and the speaker is located on the bottom of the diecast aluminum box. In a noisy environment it is sometimes hard to hear the MMS-1.

The MMS-1 will be particularly interesting to groups and clubs looking for help with code instruction. Students can sit down with the MMS-1 without a teacher's assistance, and program the unit at any level they are comfortable with.

For more information, contact Spectrum International, Inc., Box 1084, Concord, Massachusetts 01742. Reader Service Number 013.

the editors
ham radio

ergonomic comfort chair

Charvoz-Carsen announces the Charvoz Dauphin CRT Chair G1500 for the Amateur Radio, home video game, and computer enthusiast. This new ergonomic chair features five functions for those who spend long hours at play or work in a sitting position. The pneumatic finger-tip controls allow the user to adjust his seated position for maximum comfort with full freedom of movement.

These chairs are designed in Italy for beauty and West German engineered for years of trouble-free enjoyment. Seats move up and down as well as tilting forward; the backrest goes up and down and inclines gently to match your lumbar/lower back needs. The backrest also tilts automatically with your back movement. This chair features the built-in lumbar-comfort support, five point star base and enclosed back-shell for added beauty.

Five fabric colors and open or closed armrests are available. For more information, contact Pat Gusoff, Charvoz-Carsen, 5 Daniel Road East, Fairfield, New Jersey 07006; telephone 201-227-6500. Reader Service Number 084.

coded squelch test unit

Ferritronics, Inc., announces the new TU-100 Coded Squelch Test Unit. The TU-100 is a microprocessor-based instrument designed to aid technicians in testing and troubleshooting sub-audible encoder and decoder circuits.

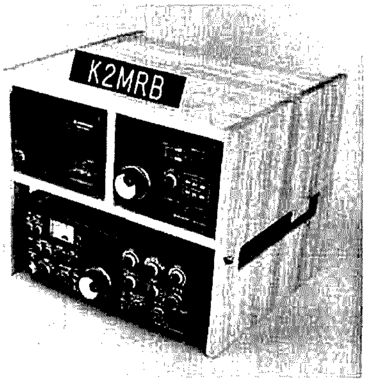
In addition to EIA CTCSS tones, the test unit works with digital codes compatible with Digital Private Line, Digital Channel Guard, Digital Quiet Channel, Digital Call Guard, etc. Used in conjunction with a monitor receiver, the test unit may be used to police shared repeaters, select unused

codes and identify unknown codes right off the air.

The TU-100 is highly portable in a durable ABS plastic case with retracting carrying handle, has a Ni-Cd battery pack, and weighs in under four pounds. For more information, contact Ferritronics, Inc., 1319 Pine Avenue, Niagara Falls, New York 14301; telephone 800-828-6884/New York: 716-282-7470. Reader Service Number 011.

Guild radio rack

New for hams is the Guild Radio Rack. The Guild Rack comes in fin-



ished solid ash. No assembly is required. Guild's radio rack comfortably holds Kenwood's TS830S/VFO230/SP230 or TS820 series, and any similar rigs. Exact measurements are: overall 16-7/8 × 14-3/4 × 14-1/2 inches, top compartments 7-1/2 × 6 inches, bottom compartment 15-5/8 × 7 inches, and it's fully vented.

The Guild Radio Rack has a suggested retail price of $59.95. For more information, contact Guild Radio Rack, 225 West Grand St., Elizabeth, New Jersey 07202; telephone 201-351-3002. Reader Service Number 086.

low-cost wind power

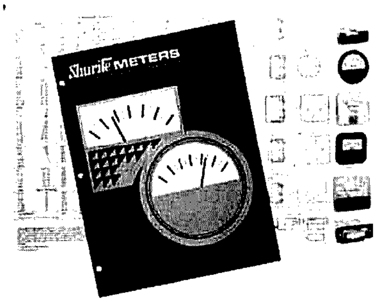
The TC25WG Helius Rotor Kit was developed as a low-cost battery charger for remote sites. Applications

NEW products

include power for camping, boats, radio operation, RVs, cottages, experimentation, and emergency situations. The patented Helius design responds to winds from any direction and is self-starting. The TC25WG rotor operates at low RPM and avoids overspeed problems common in propeller designs. Virtually no maintenance is required and the tough Lexan vanes resist sun, snow, sleet, and extreme temperatures. Assembly is easy with regular hand tools. The complete rotor kit costs $479, including detailed manuals. A free information package is available from Thermax Corporation, One Mill Street, Burlington, Vermont 05401. Reader Service Number 088.

panel meter brochure

Shurite Meters' free brochure offers a choice of more than 260 standard-range ±5 percent accuracy instruments. A standard-range chart shows range, resistance, and stock numbers in vertical columns, as grouped in five major product categories. The horizontal section of the matrix shows dc microammeters, dc milliammeters, dc voltmeters, ac milliammeters, ac ammeters, and ac voltmeters.



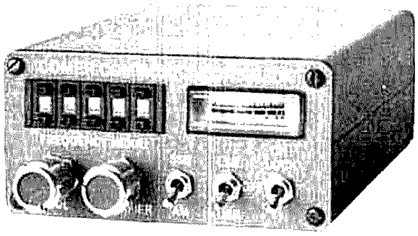
Other contents show specifications, standard options, outline drawings, custom dials and special ranges. Line voltage testers and battery testers round out this catalog.

For further information, write to

Shurite Meters, Inc., 577 Grand Avenue, P.O. Box 1848, New Haven, Connecticut 06508-1848; telephone 203-624-1188. Reader Service Number 087.

small receiver

The Comer Communications R30K is an extremely small general coverage communications receiver, measuring only 2 × 4 × 6 inches. Despite this small size, the R30K performs as well as receivers eighteen times larger. Frequency coverage is from 50



kHz to 30 MHz, and selection is by means of a five-digit push wheel switch giving 1 kHz resolution. A VFO is provided to interpolate between the 1 kHz steps.

Reception modes are AM, LSB, USB, and CW. The receiver operates from a nominal 13.8 Vdc supply and requires an external loudspeaker or headphones. A connector has been provided at the rear of the unit with outputs from the first, second, and carrier oscillators, and also with an input/output of the AGC line.

The receiver is a dual-conversion superheterodyne with a high first i-f. Frequency control is by means of a fully digital dual loop frequency synthesizer. Selectivity is obtained by six-pole crystal filters giving 2.4 kHz on SSB and 6 kHz on a-m. The use of modern high-level integrated circuit mixers give the receiver a very respectable dynamic range while holding the power consumption to a very reasonable level. Normal fast attack fast release AGC is used on a-m, while a fast attack, slow decay, with one-second hold on no signal AGC system is used on SSB. The audio output provides 5 watts at 5 percent distortion to give good results in a

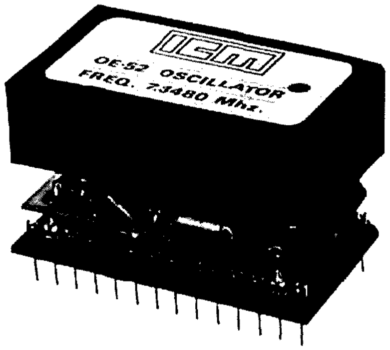
NEW products

mobile environment.

For more information, contact Comer Communications, 609 Washingtonia Drive, San Marcos, California 92069; telephone 714-744-3215. Reader Service Number 085.

precision frequency control

The latest design in International's high reliability OE series oscillators features TTL output in a 28 pin DIP package. The output of the OE-52 and OE-53 is sufficient to drive up to ten TTL loads. The TTL output offers many new general instrumentation design applications in clock circuitry microprocessing and precise frequency control.



Specifications: input voltage — +5 Vdc; temperature stability — −30 to +60 degrees C; OE-52 — ±0.0002 percent; OE-53 — ±0.0005 percent; output voltage — 5.5V PP, 2.9V RMS; trim range — ±10 ppm minimum; size — 0.430 × 1.45 × 0.795 inches.

For more information, contact International Crystal Mfg. Co., Inc., 10 North Lee, P.O. Box 26330, Oklahoma City, Oklahoma 73126; telephone 405-236-3741. Reader Service Number 012.

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contents

REFLECTIONS

Happy Birthday ham radio!

This issue, on our fifteenth anniversary, marks the completion of fifteen continuous years of providing the Amateur community with important technical information. We have attempted, and we hope succeeded, in bringing you a consistent flow of data: past techniques, still useful and important today; present techniques for solving today's problems; and looks into the future technologies that so rapidly are becoming a part of today.

This issue also signals the start of our sixteenth year of publishing in the same tradition, with articles that clearly emphasize the ham's true investigative spirit: always improving on existing techniques, making gear smaller, better, and less expensive. This month, that is best brought out by the articles by WA4ZXF and K2BT. Douglas Glenn (WA4ZXF) provides more than just another detailed design (and construction information) on a 15-meter transceiver. He shows, for example, how different operating functions can ingeniously be combined in the same circuit, thereby economizing on the number of components. A typical example is the use of the first mixer as a converter in the receive mode and as a balanced modulator during transmit.

Just because you're working with inexpensive components from the junkbox doesn't mean you have to put up with limited performance. Doug designs around them, taking full advantage of their characteristics. And the best part is that even if you don't build this unit, an understanding of a complete SSB transceiver is yours just for the reading.

Forrest Gehrke, K2BT, closely examines rf bridge operation — its inherent inaccuracies and limitations — in his article "Precision Noise Bridge." He shows, in his design, how real world impedances can be confidently and accurately measured. Forrest has used the noise bridge in building one of the most competitive 75-meter directional arrays in the world: the "4 square" (much more about that in several upcoming issues). With his bridge, he is able to accurately determine self and mutual impedances that unfortunately don't always come close to a resistive 50 ohms.

On a separate tack, Woody Smith, W6BCX, continues where he left off (in his Part 1, last month) with construction details for the half-square and Bobtail curtain. He closely examines feeding methods and locations, showing their influences on pattern and on the frequency bands that are usable with the Bobtail.

Dennis Mitchell, K8UR, provides additional insight into the GaAs FET technology. His circuit, designed around the Alpha Industries ALF1023 device, incorporates features that most hams can readily use in achieving higher-gain, low-noise preamplification *inexpensively* on 2-meters. This is especially significant since the trend, industrially, has been to use GaAs FETs at microwave frequencies (where their impedances are closer to 50 ohms), avoiding designs on the lower frequencies. Dennis evaluates seven different FETs for noise figure, gain, 1-dB compression, and cost.

Ernie Franke, WA2EWT, explores capacitively coupled hybrids, devices useful at any power level. A little bit of thinking shows several applications where a continuously variable phase shift at rf could be very useful, be it in the laboratory or out on the range (antenna range, that is).

These, as well as the other articles in this issue, are but a sampling of what we at *ham radio* have in store for you in 1983. Consider what you would like to see in the following issues; it will probably be there. Our crystal ball has recently been calibrated to accurately focus in on your needs. Nevertheless, we still welcome all of your communications. Let us know your views!

While on the subject of what *ham radio* has in store for the upcoming year, we take this opportunity to thank the people who've made *ham radio*'s continued existence possible. First, of course, are our readers, without whose interest in excellence in radio communications there would be *no ham radio* magazine. We thank our advertisers, who have chosen our magazine for presenting you with the latest in communications gear — be it transceivers, antennas, or accessories. They are worthy of your support, for it's all a complete cycle — readers — advertisers — *ham radio* magazine. Let the advertisers know your feelings and needs. Tell them "you saw it in *ham radio*."

We especially wish to thank our charter advertisers — see the list below — identified with the symbol on their ads throughout the magazine. Readers who have the complete collection of *ham radio* magazines (doesn't everyone?) will recognize these companies as having been with us in the beginning.

Without further ado, as my British friends are wont to say: "Lets get on with it."

Rich Rosen, K2RR
associate publisher/technical editor

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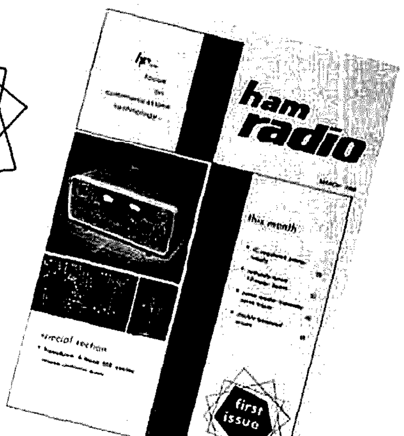
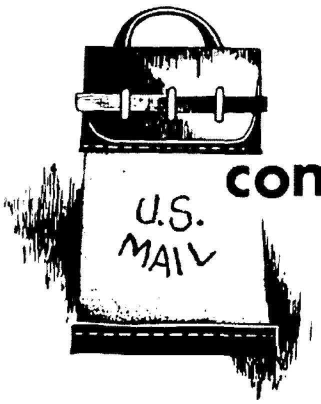




comments

debunking myths

Dear HR:

The vertical versus dipole comparison by Bill Orr in "Ham Radio Techniques" (*ham radio*, October, 1982) should help debunk the myth that a quarter-wave vertical over a good radial system is automatically a low-angle radiator. An earlier article by John Belrose (*ham radio*, September, 1981, page 36) first got me wondering whether verticals are automatic guarantees of a DX edge over dipoles, inverted vees, and other simple antennas. Belrose showed that a quarter-wave vertical over very poor ground but with an excellent ground system has a radiation angle of 30-35 degrees, about the same as a half-wave dipole erected a half wave above ground. This has apparently been confirmed by Orr in his extensive 10 MHz listening comparisons.

What makes or breaks a typical vertical's low angle of radiation is the conductivity of the ground, out to 50 or more wave lengths from the antenna. (This factor has a much smaller influence on the take-off angle of a dipole of a half-wavelength or more above ground.)

Where does this leave the vertical which has become so popular for high frequency communications? I'd say that for 160 meters, 80 meter, and 40 meter long-haul QSO's, the vertical is still the best of the simple antennas for the average ham. A dipole must be a half wavelength above ground to give the 30 degree take-off angle the vertical will give at ground level. Most city-dweller hams can't easily get a dipole that high. However, as Bill Orr demonstrates, the dipole probably becomes the better simple antenna from 30 meters on up. Exceptions would be the ham on a boat, in a salt-marsh, or possibly in an agricultural setting where continual use of fertilizers has given the soil superior conductivity out to 50-100 wavelengths from the antenna. Verticals in these settings will have take-off angles of 10-15 degrees. For Amateurs fortunate to be operating in these settings, the vertical still reigns supreme among the simple antennas for DXing from 30 meters on up.

Mark Bacon, WB9VWA
Decatur, Illinois

the 432 Yagi is alive and well

Dear HR:

The excellent article, "Requirements and Recommendations for 70-cm EME," by Joe Reisert (*ham radio*, June, 1982) mistakenly notes that "the K2RIW (432 Yagi) has gone out of production."

The K2RIW Yagi, the original RIW 432-19, is still in production by RIW Products. In fact, to assist the home-brewer, kits for the 432-19 are also available that include all the hard-to-get parts at a modest price.

George A. Flanagan, W2KRM
RIW Products
Box 191
Babylon, New York

presstop de W9JUV

BOTH THE "NO-CODE" LICENSE AND AMATEUR EXAM ADMINISTRATION PROPOSALS are out, after the Commissioners voted to adopt the appropriate Notices of Proposed Rule Making on January 20. The "No-Code" is PR Docket 83-28, and the exam proposal is PR Docket 83-27. Both will have comment due dates of April 29. While the texts of both these crucial NPRMs had still not been released at press time, much of their content was already available.

The "No-Code" Proposal Is Very Much As Predicted in last month's Presstop. It offers two alternatives. The first is simply removing the Element 1A (CW) requirement from the present Technician Class exam, granting the successful applicant full privileges above 50 MHz. If such a licensee later wished to operate below 50 MHz, he could then be granted access to Novice frequencies after successfully demonstrating his CW ability to a higher-class licensee. A record of that "certification" would have to be kept in his station records, but there would be no notification to the Commission.

The Second Alternative Would Be The Establishment of a new "Digital Operator" class of license. As proposed, the "Digital Operator" would have full access to all Amateur bands above 144 MHz. Commissioner Dawson expressed a great deal of concern that restricting the frequencies available for this license would make it too unattractive. A new exam, "Element 5," perhaps incorporating present exam elements, would be written for the "Digital Class."

The FCC's Decision To Go Ahead On The "No-Code" License was made in spite of heavy opposition by the ARRL, and the item actually received relatively little discussion at the agenda meeting. League President Vic Clark, W4KFC, had urged the Commissioners, both in personal letters and visits, to hold off on no-code for at least 18 months to give the Amateur-administered exam program a chance to get rolling, but the Commissioners have been quite interested in seeing a no-code license established and were not receptive to seeing it delayed at this time. It's likely the League's position is not set in stone, either. In a Westlink interview, League Counsel Chris Imlay, N3AKD, noted that neither membership nor director opposition to no-code was unanimous and that the League position would certainly be up for review at the April Directors' meeting. In view of the strong sentiment in favor of no-code apparent within the Commission, it seems likely the League's final position could be one of tempering the no-code license to make it more acceptable to the majority of its members rather than continuing to oppose no-code outright.

The Proposal To Set Up A Volunteer-Administered Licensing Program generated much more discussion among the Commissioners and staff than did no-code. A good deal of the NPRM came from the ARRL's Petition for Rule Making, but with many changes from both within the Commission and from comments filed by the QCWA and others. One concern was that there must be more than only one Volunteer Examination Coordinator group; QCWA, IEEE, and OMICK (a national group of black Amateurs) were all mentioned as other possibilities along with the League. Another concern was that membership in a given group should not be required for examiner certification.

Specifics Of The Exam Proposal include the use of three-person exam teams, with the team chief being an Extra Class Amateur and team members either Advanced or Extra. CW exams for General or above would be administered by an Extra Class, and exams for the Extra Class license would require an all-Extra team. Exam questions would be drawn from lists made up by the FCC from submissions by individuals or groups. Teams could issue Interim Permits, just as Field Offices do now. Team members would have to be over 18, and could not be related to the applicant. No one working for a manufacturer or distributor of Amateur equipment, or for a publisher of training material, could serve as an examiner.

Though The ARRL Had Pushed Hard To Have The Novice Exam Included in this new exam proposal, the Commissioners elected to leave it as proposed late last year in PR Docket 82-727. Under the simpler procedure proposed in that NPRM, a Novice applicant could receive his license a few weeks after taking the exam. Under the more cumbersome procedures of the proposed higher-class exam program, Novice license processing would be considerably slower.

THE "AMTOR" DIGITAL SYSTEM WAS APPROVED for general Amateur use on 80 through 10 meters by the FCC on January 27. AMTOR is a synchronous RTTY technique, similar to the marine service's SITOR, which verifies message content and accuracy, thus permitting communications with much lower signal levels than RTTY otherwise requires. A number of Amateur stations have been using AMTOR under FCC Special Temporary Authority, and W1AW should begin using it within a few weeks.

GE'S PROPOSAL FOR A NEW 900-MHZ PERSONAL RADIO SERVICE WAS ADOPTED as an NPRM by the FCC on January 20. Located just below 900 and above 937 MHz, it would use self-identifying transmitters, andhave automatic access to telephone service via the user's own base station or commercially operated repeaters. A similar service, though without repeaters, has just begun operation in Japan at 903-905 MHz (Region 3 did not get the 902-928 MHz Amateur band).

This Proposal Is Significant To The Amateur Service for several reasons. First, as a much more useful version of CB it's very likely to attract the interest of sophisticated would-be users who would otherwise be good candidates for Amateur Radio, with or without a no-code license. Second, the technology that such a service could support would spill over into Amateur Radio, particularly for the new 902-928 MHz band.

Comments On The Proposed New Service Will Also Be Due at the FCC on April 29. It has been assigned General Docket 83-26 by the FCC.

15-meter sideband transceiver

A construction project of solid design with step-by-step explanations of each circuit function

With some careful shopping, it's possible to build a 15-meter sideband transceiver for less than $100. The design presented here is for a basic rig at minimum cost, but with excellent features:

Receiver: 0.5 μV sensitivity; 1.0 μV agc threshold; 0.5-watt audio output; 2000-Hz selectivity; agc; 80 dB compressed to 4 dB; CW audio filter.
Transmitter: 10 watts PEP; fully adjustable output; CW break-in keying; 50-dB carrier and LSB suppression.

I designed this rig for 15-meters because propagation conditions on this band put low-power operators at less of a disadvantage than on the other bands. If you prefer, the rig could be built for the 20-meter band by using a different offset crystal frequency and by adjusting several tuned circuits. With an output that is fully adjustable from a pot, it makes a good exciter for a linear amplifier.

theory of operation

Some of the circuit details may appear unfamiliar or unusual, but I will provide enough of an explanation for you to understand how the circuit operates. Component substitutions may be made to hold the cost down.

Let's start by reviewing the receiver signal path, beginning at the antenna jack (see **figs. 1** and **7**). No antenna relay is used. Instead, the input signal is coupled from the antenna low-pass filter to the receiver rf amplifier through a series-resonant circuit, C1-L1. This series-resonant circuit is used both as a T-R switch during transmit and as an rf attenuator during receive.

A series L-C circuit presents a low impedance at resonance, while the center junction is at a high impedance. If you put a variable resistor from this junction to ground, you create a variable attenuator. But instead of a resistor, use the collector impedance of a transistor. The transistor looks like a small capacitor (junction capacitance) in parallel with a variable resistance. The resistance value is controlled by the dc base current, as shown in **fig. 2**, as long as the signal amplitude doesn't forward-bias the transistor junctions. A typical value is less than 10 ohms with 1 mA of base current, and it increases to nearly infinity as base current reduces to zero. The effect of the junction capacitance must be compensated for, so avoid transistors with large capacitance. A diode across the transistor protects the receiver's front-end from other transmitters feeding antennas on the same tower. The circuit can also be used for filtering if desired, but it doesn't have to be. In this case, filtering was not a concern.

There is a trade-off that must be made. As the L/C ratio gets larger, you get higher Q and more attenuation range. But as the capacitor value approaches the value of the shunt capacitance from the transistor and diode, the capacitive voltage-divider effect significantly attenuates the input signal. The L1-C1 values were chosen for a good compromise, and they have low Q so that fixed components with normal tolerances can be used.

The input signal from the rf attenuator is coupled to an rf amplifier, Q2, with about 20 dB of gain. Tuned transformers are used at the input and output of this stage to prevent image and i-f feedthrough. The transistor that I used was an inexpensive general-purpose type, the 2N3904. This transistor's input impedance is about 50 ohms, so the transformer has unity impe-

By Douglas Glenn, WA4ZXF, 715 Fairlane Drive, Lewisburg, Tennessee 37091

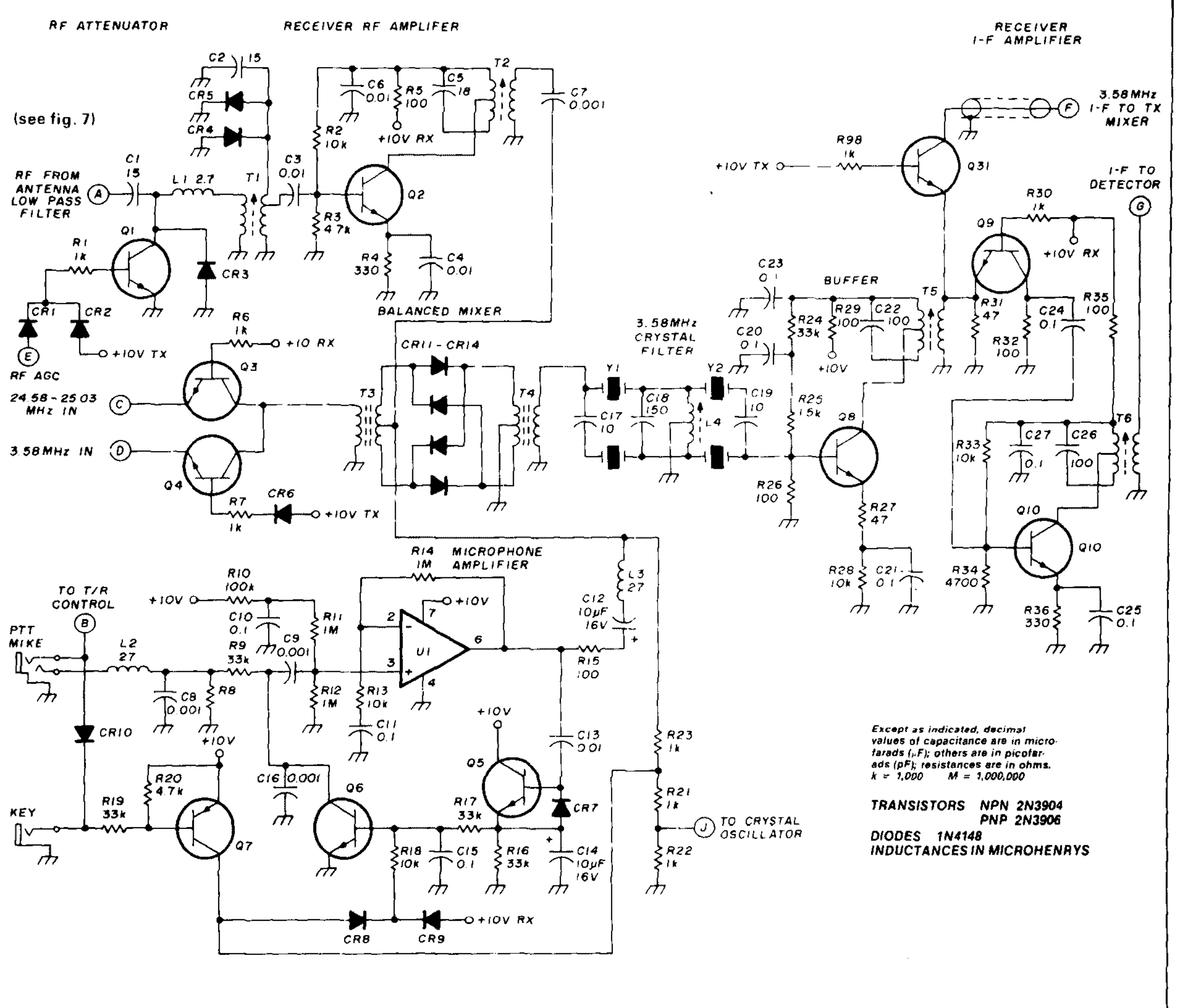


fig. 1. The heart of the rig is a homebrew balanced mixer and crystal filter. Schottky diodes in the balanced mixer would be good substitutes.

dance ratio from input to base. The circuit *Q* is not allowed to get too high, which would hurt sensitivity at the band edges. Diodes across the high impedance winding give extra protection from high-level signals.

The output of the rf amplifier is coupled to the first mixer by a capacitor, C7. This same mixer is used as the balanced modulator during transmit. The mixer is a double-balanced, diode-ring-type, selected because it provides good performance, is inexpensive, and is easy to build. The output transformer is wound to provide the impedance needed by the crystal filter instead of the usual 50 ohms. Good carrier suppression requires matching of the diodes. But this is easy, as explained later.

The mixer input that is used for the signal from the receiver's rf amplifier is also used as the audio input during transmit. To avoid the need for a switch of some kind at this mixer input, the high frequency signal in receive is coupled through capacitor C7, blocking the low-frequency audio during transmit. Similarly, coil L3 in series with the audio blocks the rf.

During receive, the local oscillator input to the mixer is from the high frequency VFO, but in transmit this same input is from a lower-frequency crystal oscillator. Two transistors (Q3 and Q4) provide the mixer input switching function. A relay could be used but it would probably cost more, take up more board space, and draw more power-supply current. This analog switch doesn't provide the high isolation of a relay, so you may notice a birdie 30 kHz above the top edge of the band (6 × 3.58 MHz) in receive. To minimize this, a diode is placed in series with the base of Q4, increasing the turn-on threshold.

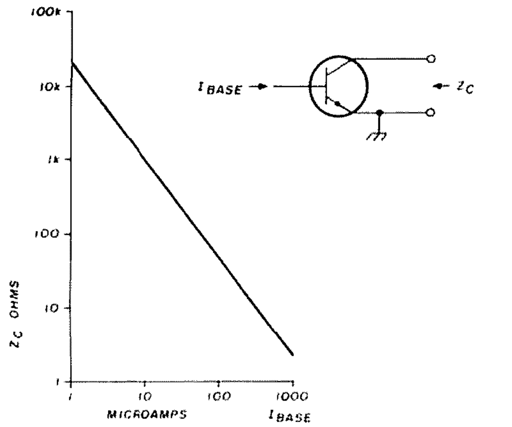


fig. 2. Resistive part of small-signal collector impedance versus dc base current. At high frequencies, the collector junction capacitance will appear in parallel with the resistance.

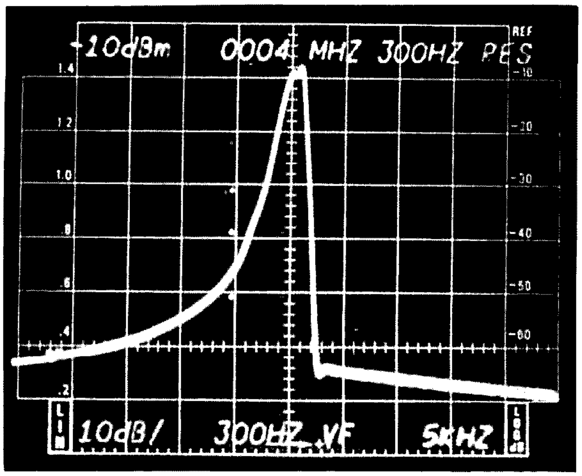


fig. 3. Crystal filter response with 1000 ohm source and load. Insertion loss is 5 dB.

crystal filter

An inexpensive crystal filter requires inexpensive crystals. At present, the most readily available and inexpensive crystals are TV color-burst crystals at 3.58 MHz, which are used here. The circuit is a standard four-crystal lattice. This requires two sets of two crystals with a small frequency difference. Since color-burst crystals all come at the same frequency, small capacitors are inserted in series with the crystals on opposite corners of the lattice to provide the needed frequency difference. The value of the capacitors establishes the filter's bandwidth. I used capacitor values of 10 pF, which results in about a 2-kHz bandwidth. Smaller capacitor values will give wider bandwidth. You should not try to make it too

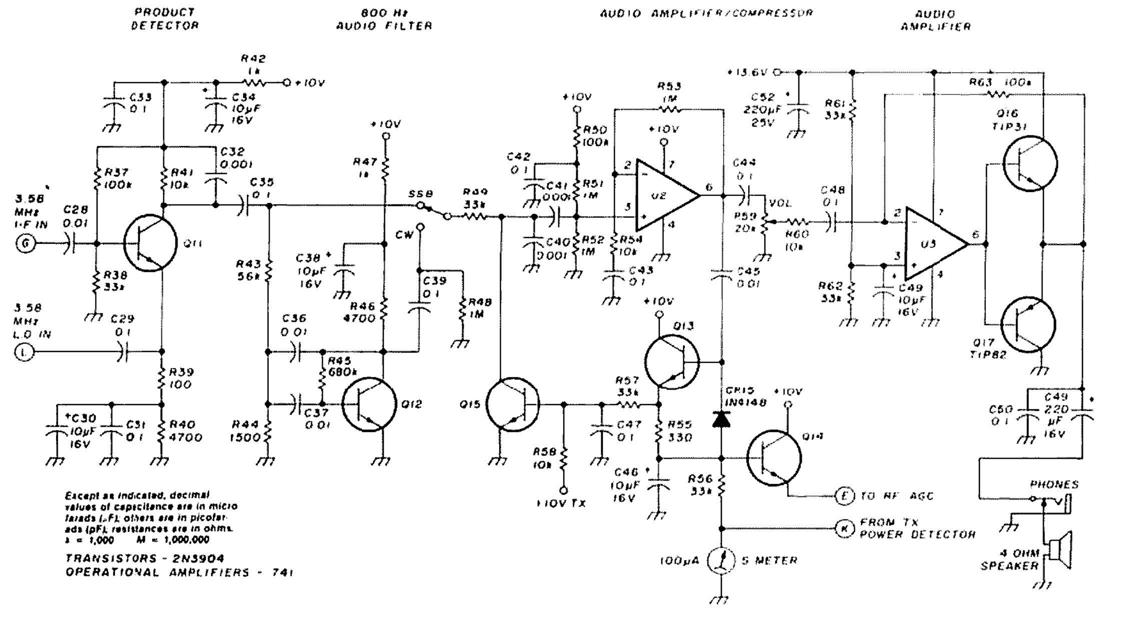


fig. 4. The audio circuit provides a lot of the receiver's gain. The input to the detector at the compression threshold is about −55 dBm.

much wider or you will have trouble adjusting the 3.58-MHz oscillator later. I used the small, round, axial-leaded, titanium-dioxide capacitors found in many TVs and fm tuners, but only because I had some handy; other ceramic or mica 5-percent-tolerance types will work. **Fig. 3** shows the shape of the filter. The shape obviously dictates use for the lower sideband. Since the upper sideband is commonly used on 15-meters, high side injection must be used in a mixer to invert the sidebands.

The filter's output is terminated by a buffer amplifier, Q8. The 1000-ohm filter load is supplied by the base bias resistors in parallel with the transistor's input impedance. Part of the amplifier's emitter resistance is un-bypassed to keep the transistor's impedance high. This sacrifices some gain, but allows for accurate control of the filter's termination impedance by the fixed resistors.

The output of the buffer amplifier is coupled to an i-f amplifier, Q10, in receive, or to the second conversion mixer in transmit. Transistors Q9 and Q31 are again used as analog switches. In receive, the i-f signal is amplified and sent to the receiver's detector. The receiver i-f amplifier is turned off during transmit to prevent the possibility that the oscillator signal might bypass the crystal filter.

receiver audio circuits

The receiver's detector (see **fig. 4**) is a single transistor, Q11, configured as a mixer for product detection. Signals feed the base while the local oscillator is injected at the emitter. The collector load of the detector is bypassed for rf, leaving only the recovered audio.

The audio is coupled to an active unity gain bandpass filter, Q12, centered at 800 Hz. Since neither high gain nor high Q is needed, I couldn't justify using an op-amp when a single transistor would do. The formulas that you would use with an op-amp don't work with the low-input impedance of the transistor. The filter Q and center frequency will be less than predicted. As shown, the filter is centered at 800 Hz with a 300-Hz bandwidth. There's no need for precision components in the filter because the filter's performance can be easily adjusted using the two voltage divider resistors at the input (R43 and R44). The small resistor to ground, R44, will vary the center frequency (less resistance for higher frequency) and the large resistor in series with the input, R43, is then changed to set the gain again to unity. A SPDT switch bypasses the audio filter for phone reception.

The wide or narrow audio from the bandwidth switch is coupled to the receiver's audio amplifier/compressor. This circuit is configured as a current-controlled *attenuator* followed by a 40-dB gain *amplifier*. I used 741-type op-amps because they are inexpensive and readily available. The op-amp is not being used near its maximum gain or slew rate, so you can get by with a 10-volt supply. The attenuator is ahead of, not inside, the feedback loop and consequently loop gain remains high at all input levels. This arrangement provides tight control of the audio level. You'll never have to dive for the gain control or claw at your headphones when a strong signal comes on frequency.

The attenuator is a voltage divider that consists of a fixed resistor, R49, and the variable impedance of Q15, which has the characteristic shown in **fig. 2**. This controls the signal level at the op-amp input. A transistor peak detector, Q13, at the amplifier's output provides drive to the attenuator in the form of base current to Q15. The *compression threshold* occurs when the peak-to-peak output of the amplifier exceeds the base-emitter diode drop of Q13. This corresponds to 210 mV rms for a silicon transistor, or 2 mV at the compressor input. The circuit features a fast-attack, slow-decay characteristic compatible with speech signals. The circuit also limits sharp noise pulses. A small resistor, R55, is placed between the emitter of the peak detector transistor and the large storage capacitor, C46. Short noise pulses are coupled directly to the attenuator, but only slightly affect the charge on the storage capacitor. A smaller capacitor, C47, is placed across the base-input of the attenuator transistor; it serves only to eliminate the audible pop that would occur when the input signal crosses the compression threshold. Once in compression, the voltage across this capacitor is fixed at the potential of a base-emitter diode drop and will not limit the response to the noise pulses or the fast attack characteristic. During transmit, the receiver's audio gain is killed by forcing the attenuator transistor to saturate. This doesn't charge the storage capacitor, so the receiver is at full gain and ready to go as soon as the transmit bias is removed.

The voltage on the storage capacitor is fed to the receiver's front-end attenuator as rf agc. An emitter follower, Q14, is used to reduce the loading on the storage capacitor and a diode is placed in series with the emitter follower output to protect it from reverse bias damage during transmit. The result is that the rf agc voltage is reduced by two diode drops. This provides delay in the activation of the rf attenuator so that the input is not attenuated until the signal is well above the noise.

The voltage on the storage capacitor is also used to operate the receiver's S-meter. The S-meter current helps to control the decay time of the receiver agc and it ensures that the capacitor discharges completely when there is no signal received. If you don't use a meter, replace it with a short; don't omit the discharge resistor. I used a surplus meter that I

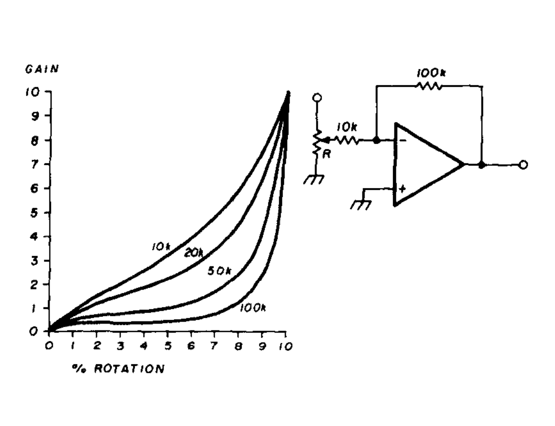


fig. 5. Gain versus pot rotation. You can simulate a logarithmic response by selecting the pot's resistance.

picked up at a hamfest for half a dollar. I painted over the face and drew a new scale. The meter doesn't read until the received signal is above the agc threshold at 1 microvolt. This corresponds to an S-level of about 3.5, so the first mark on the meter is S-4. You can add capacitance across the meter's terminals to increase the damping of your meter if you want. You'll have to *experiment* to find the right value.

The audio output amplifier uses an op-amp followed by a high current buffer. The buffer transistors are not biased into the linear range, but they are included inside the feedback loop. The slight amount of crossover distortion won't be noticed under normal conditions, and it's a small price to pay for circuit simplicity and the saving in quiescent bias current. A capacitor was found to be necessary across the output to bypass rf pickup that had the effect of making the audio amplifier squeal during transmit, especially when I was using an external speaker.

The output amplifier is operated directly from the external 13.6-volt supply. Power supply hum is not a problem because of the op-amp's good power supply rejection ratio and a heavily bypassed reference (noninverting) input. Operating from a higher supply voltage also increases the maximum output power available. In addition, stability of the audio circuits is improved because the supply voltage for this high level circuit is isolated from the supply for the low-level circuits.

One interesting effect of this circuit is that a linear volume control gives a non-linear response to the cir-

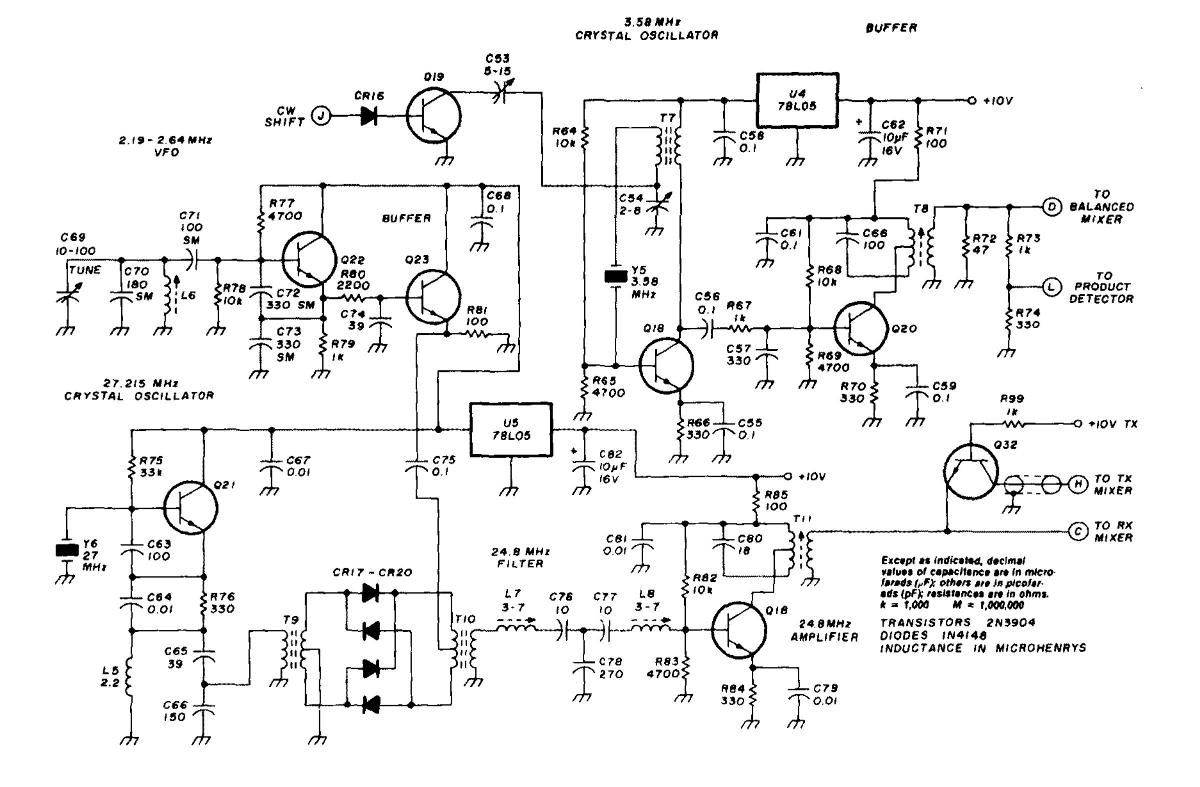


fig. 6. All of the oscillators are supplied by 5 volt, 3-terminal regulators for stability.

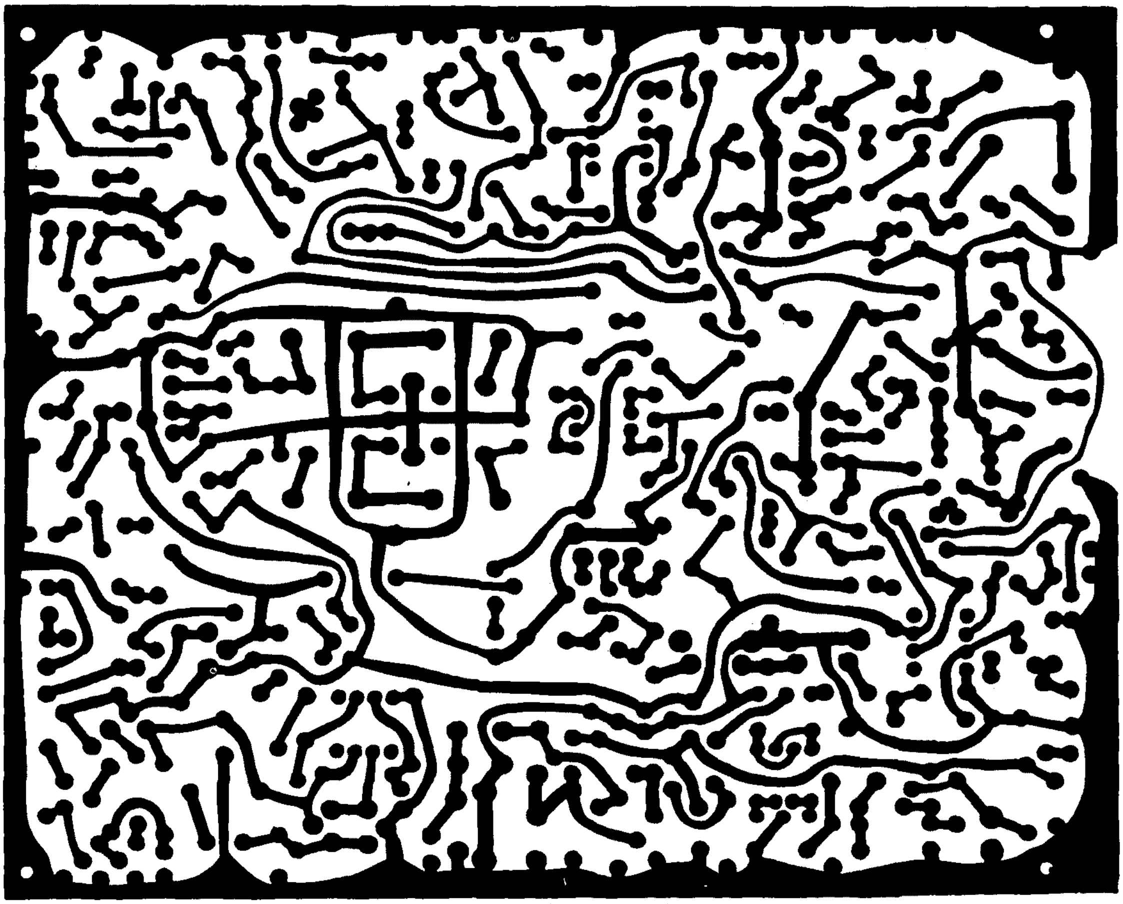
cuit. The shape is not the true logarithmic curve of audio controls, but the overall effect is similar. Part of the pot's resistance adds to the resistance in series with the op-amp input and so changes the circuit's gain slope as the pot is rotated (see **fig. 5**). The effect is more pronounced for pots with larger resistance values. If you use a pot that is significantly more than 20 kilohms, you may want to proportionately increase the input and feedback resistors.

transmitter circuits

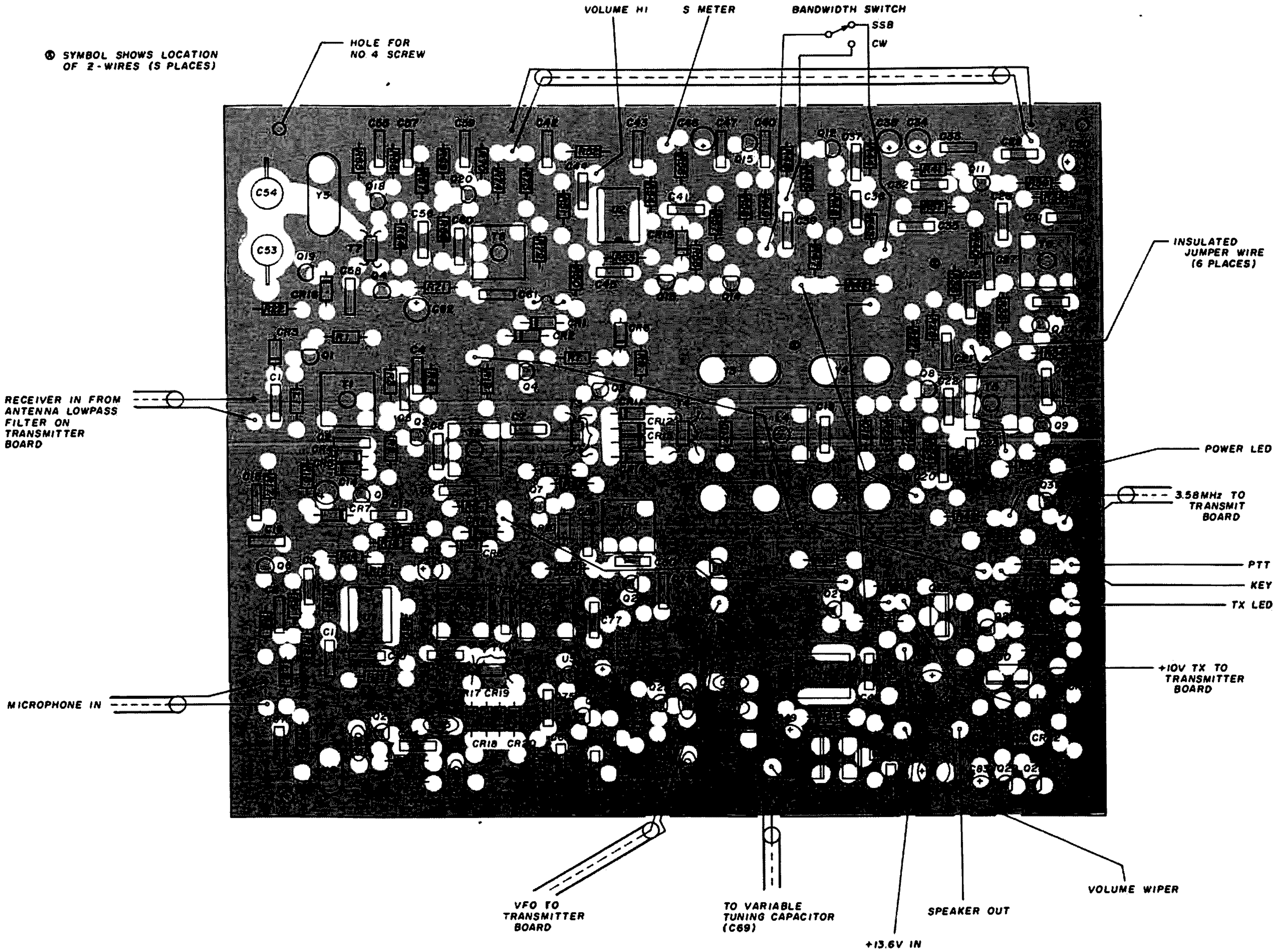
Transmitter circuit review begins at the microphone input. (see **fig. 1**) The microphone amplifier uses a compressor amplifier (U1, Q5, Q6, etc.) nearly identical to that used in the receiver. By using a compressor that maintains a constant audio level, the other circuits can be optimized for this level. The only circuit difference (from the receiver's compressor) is one missing resistor. The resistor is omitted since the microphone amplifier doesn't have to contend with receiver impulse noise.

The amplifier input is terminated for a dynamic microphone. This is because the inexpensive, readily available microphones for citizen band sets are usually of this type. A filter consisting of a coil and capacitor keeps rf pickup out of the radio. A higher input impedance, for another type of microphone, would require some redesign of the input.

The output of the microphone amplifier is coupled to the balanced mixer. The carrier signal is suppressed at least 30 dB by the balanced mixer. The signal then passes through the crystal filter, which removes the upper sideband and suppresses the carrier another 20 dB. When keyed for CW, the mixer balance is upset intentionally and the carrier comes up to full level. A transistor switch, Q7, supplies current to the dc-coupled input of the mixer to unbalance the mixer. At the same time, this switch sends current to the current-controlled attenuator of the microphone amplifier. This kills the gain of the microphone amplifier and prevents audio from being superimposed on the CW signal.



Main circuit board PC layout.



Main circuit board, groundplane side with components layout superimposed.

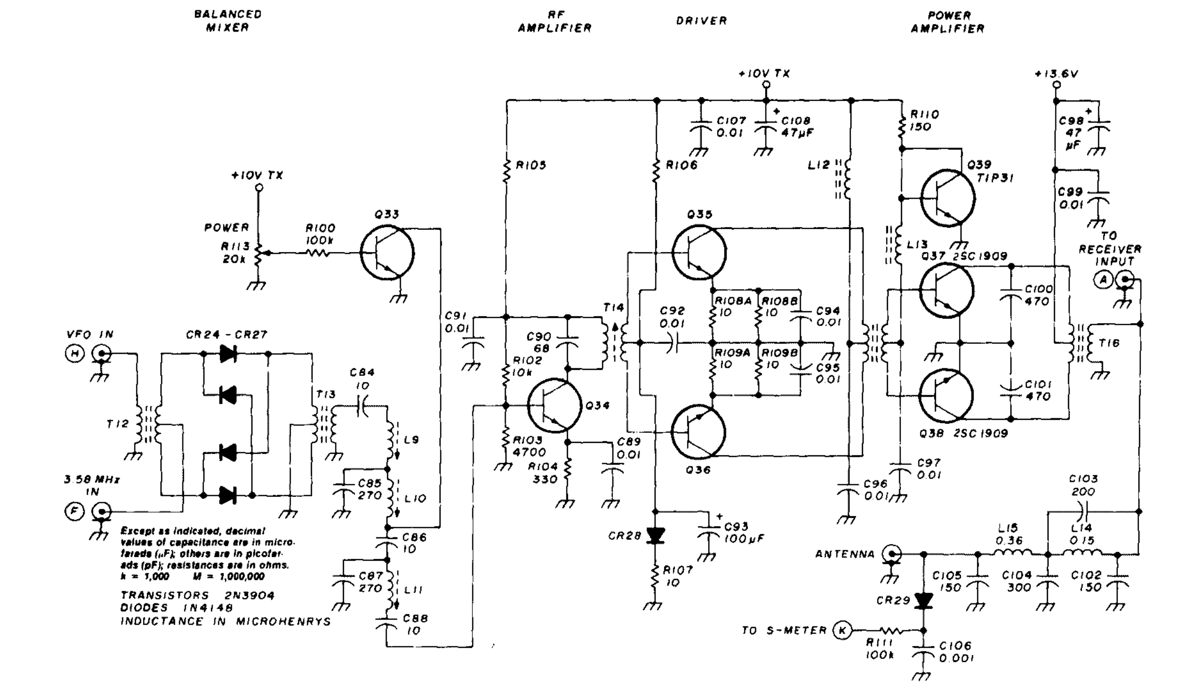


fig. 7. The power amplifier draws nearly 2 amperes at maximum output, but idle current is quite low; less than 20 mA.

carrier oscillator

The carrier oscillator (see **fig. 6**) also uses a 3.58-MHz color-burst crystal. The oscillator circuit configuration was selected for its very good frequency stability. The phase inversion, from collector to base of the transistor, is provided by a transformer rather than some form of resonant circuit. This helps stability by removing all frequency-sensitive components except the crystal and a small capacitor. The capacitor shifts the series resonance of the crystal-capacitor combination to the desired frequency. The transformer is constructed by winding magnet wire on a ferrite bead. The capacitor can be a fixed type that's selected for the proper frequency, or a variable if it's a stable type.

The capacitance is located in the ground leg of the series circuit so that another capacitor can be switched in parallel to shift the frequency. This is done when the rig is keyed for CW. The purpose is to shift the frequency by 800 Hz. When properly adjusted, the output in transmit will be the same frequency as a received signal centered in the CW audio filter passband. A transistor is used as the switch. A low capacitance transistor is absolutely necessary in this location.

The output of the oscillator is coupled to a buffer stage, Q29, through a low-pass R-C section. The R-C section does three things: The series resistor isolates the oscillator from the rest of the circuit, the low-pass section filters the signal providing a sine-wave output, and the signal is adjusted for optimum drive to the mixer. The value of the series resistor, R66, or shunt capacitor, C57, can be changed, if necessary, for more or less mixer drive. The best drive level would be +7 dBm (a 0.5 V rms). However, this level is too high for the transistor analog switch. I set the drive level to approximately 0.35 V rms (+4 dBm) with satisfactory performance. A voltage divider across the output of the buffer sets the proper drive level for the receiver's detector, about 60 mV. This level is not critical. Any value from 40 to 220 mV will work, but you should keep it toward the low end of the range. At the higher drive levels, the receiver's detector is more susceptible to stray signals such as the ever-present 60 Hz hum.

variable frequency oscillator

The high-frequency VFO that tunes the transceiver is critical to the rig's performance. It should be stable and not change frequency when the transmitter is

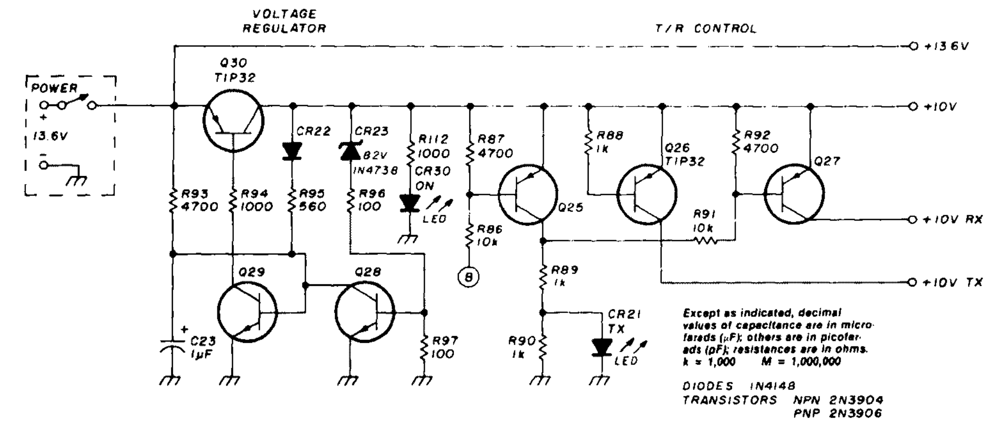


fig. 8. The voltage regulator and T/R control provide dc power to the various parts of the circuit.

keyed. The design uses a low-frequency variable oscillator which is mixed with a high-frequency crystal offset oscillator to arrive at the needed injection frequency.

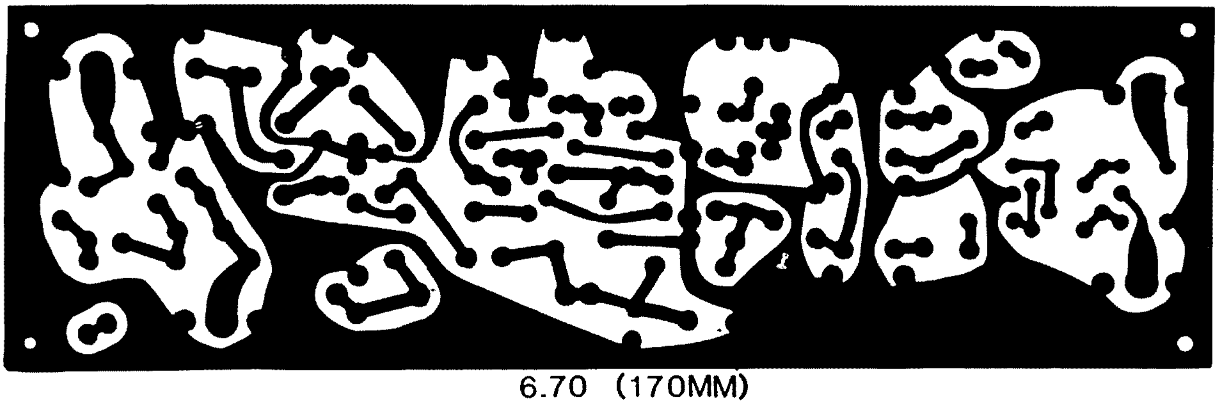
High-side injection is necessary to invert the sideband, as explained above. To cover 15 meters, the injection frequency is 24.58-25.03 MHz. The least expensive and most readily available crystals, after color-burst crystals, are 27 MHz units intended for multichannel citizen band HTs. One of these is used for the offset oscillator. They are usually sold in T-R pairs for less than $5. I used a channel 21 transmit (27.215 MHz) crystal. The exact frequency is not very important since the VFO is aligned to set the final frequency. If you have a choice, use a higher channel crystal since this will improve spurious responses. The output of the crystal oscillator is coupled to the offset mixer by a capacitive voltage divider that couples the proper level to the mixer.

The low frequency variable oscillator, Q22, is a standard Seiler type that tunes 2.19-2.64 MHz to cover 15-meters. Silver mica capacitors are used in the frequency determining circuit for improved stability. The variable capacitor used is from the oscillator stage of a two-section unit salvaged from a five-tube ac-dc broadcast receiver. If you use a variable capacitor that is significantly different, you may have to size some of the fixed capacitors proportionately. This is particularly true if your variable has more capacitance, because the oscillator quits if the L/C ratio gets too low. Increasing the value of C71 usually cures this problem. A 5-volt, three-terminal regulator, U5, operated from the regulated 10-volt supply, provides a stable, doubly regulated supply that is also used by the offset crystal oscillator. Frequency stability is good. Don't use just a zener diode; use the IC regulator. The variable oscillator output is coupled to the offset mixer through an R-C section and emitter follower, Q23, that isolates the oscillator, reduces signal harmonics, and sets the signal to the proper level for the mixer.

The mixer is a doubly balanced, diode-ring type with a two-pole, series-resonant type filter at its output. By using a balanced mixer, the oscillator feedthrough at 27 MHz is suppressed reducing the following filter requirements. The mixer design is for a nominal 50 ohms at each port, and a commercial mixer can be substituted here. The spurious responses that result from the undesired mixer outputs, as well as the transceiver image, are above the 15-meter band so that the antenna low-pass filter aids the resonant circuits of the transceiver in minimizing spurious signals. A buffer amplifier, Q24, follows the filter that provides +3 to +4 dBm drive level to the mixers.

transmitter rf circuits

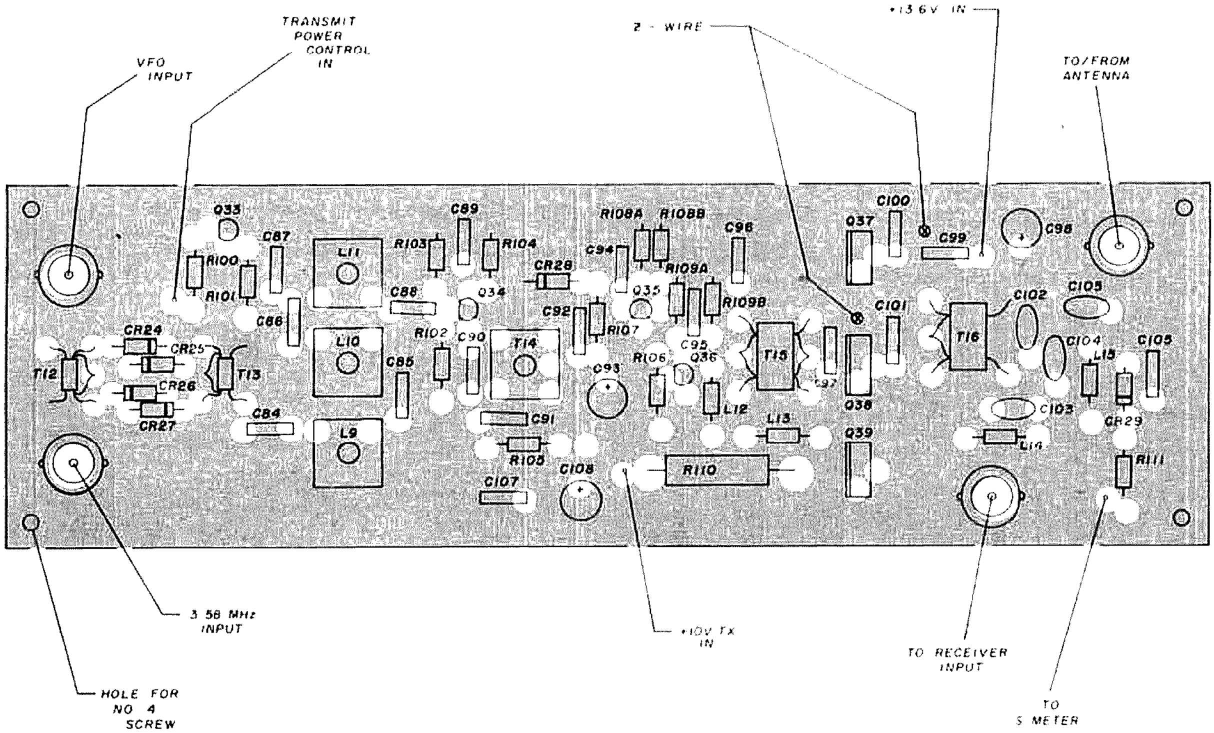
The rf circuits of the transmitter (**fig. 7**) are activated by the microphone PTT or key. The modulated 3.58-MHz signal is applied to a balanced mixer where it mixes with the variable frequency oscillator injected signal. The balanced mixer is the same type as the one that mixes the two oscillator signals and is at the normal 50-ohm impedance level. (A commercial mixer could be used instead.) The mixer output is passed through three poles of series-resonant filtering. This is the same type of series-resonant circuit used elsewhere. A shunt transistor, Q33, is used at the second pole. The base bias of this transistor is controlled by a pot and it becomes the transmitter power control. The fixed resistor in series with the transistor's collector determines the maximum power reduction. The control can vary the transmitter output at least 30 dB (10 watts to 10 milliwatts). The



Transmitter board PC layout.

attenuator is placed after all of the mixing stages to attenuate the low level spurious signals along with the desired output.

The mixer output, after filtering, is amplified in turn by a class A stage and a push-pull driver stage running class AB, prior to the final. The final amplifier, also running class AB push-pull, uses a pair of transistors intended for the output stages of 5 watt citizen band transmitters (2SC1909 from Radio Shack). Base bias current is controlled by transistor Q39 connected as a current mirror. The bias transistor is a general purpose TO-220 type that is attached to the same heatsink as the finals for temperature compensation. The input and output transformers of the final are wound on the two-hole ferrite balun cores that are found at the VHF input of many TV sets. The ones I used were salvaged from the coupler of a TV game. The final output is passed through a low-pass filter. The filter was designed to put a notch at the mixer image frequency, which is helpful in



Transmitter circuit board, groundplane side with components layout superimposed.

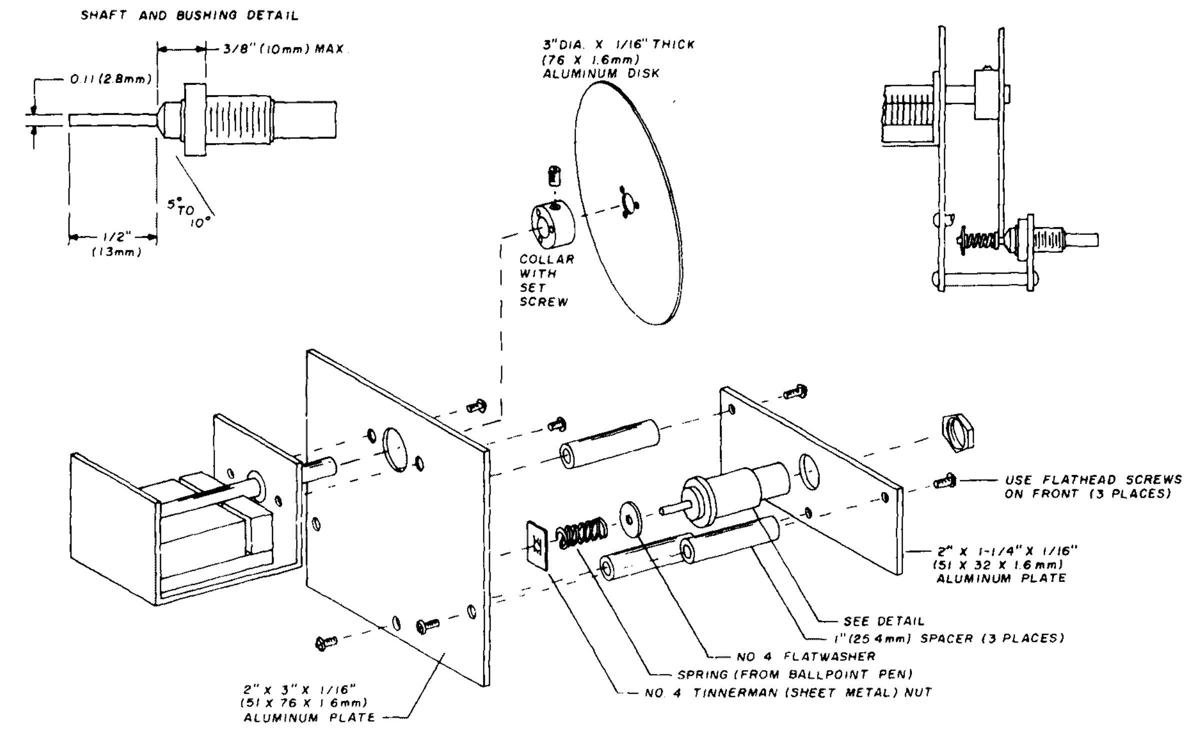


fig. 9. Exploded view of the 24:1 slow-motion drive, with the shaft and bushing detail at the upper left, and a side view of the assembly sequence at the upper right. One of the flathead screws on the front goes through the front panel to prevent rotation of the assembly around the bushing.

both transmit and receive. A diode detector at the antenna output supplies a readout of transmitter power on the front panel S-meter.

power and control circuits

The dc power for all circuits, except the transmitter final and the receiver audio output (**fig. 8**), is supplied by a 10-volt regulator. The design, which was adapted from one in a Fairchild Semiconductor application note, provides good regulation even when the input voltage drops close to the regulated level. The schematic shows an 8.2-volt zener diode, but a 7.5-volt or 9.1-volt diode also works if R96 is adjusted to compensate. Don't use R97 to adjust the voltage; its purpose is to set the zener current. The circuit also has a coarse form of current limiting. This occurs when Q29 saturates and the pass transistor, Q30, gets no more base drive. The point of current limiting can be set by adjusting the value of R94.

The dc power is switched to the various sections of the transceiver by a T-R control circuit. When activated by the PTT switch or CW key, power is applied to the transmitter circuits and removed from some receiver circuits. Current is also supplied to the several transistors that act as variable impedances throughout the transceiver. Most of the receiver circuits do not have the dc power removed in transmit. This enables quick recovery of the receiver for break-in keying.

construction

All of the circuits are contained on two doubled-sided printed circuit boards. The transmitter's mixer and amplifier chain are on one board, and everything else is on the second. This separates the high-level transmitter circuits from the other low-level circuits. You'll notice from the board layouts that the top (component) side is devoted to ground plane. Jumper wires are used where necessary rather than disturb the continuity of the ground plane. If, like me, you don't have the capability of plated-through holes on circuit boards, solder a number of component leads on top and bottom to link the ground plane. I usually did this only on resistor leads and the few Z-wires where access to the top of the lead is easy. This is particularly important on the transmitter board and especially for the emitter leads of the P.A. output transistors which need a good solid ground.

The transceiver was housed in a cabinet (Radio Shack #270-270) that provides plenty of space. The main PC board is mounted on the bottom of the cabinet using metal standoffs, which grounds the circuit

to the cabinet. The transmitter board is mounted on the rear wall the same way. The layout of the transmitter board includes provisions for four PC mount phono jacks, that are used for the rf connections. This was done to simplify removal of the board, but a soldered connection is the obvious way to go if you can't find PC phono jacks. The heatsink for the transmitter's final is a 2.5 × 4 inch (60 × 100 mm) piece of 1/8-inch (3-mm) aluminum plate that is secured to and insulated from the three power transistors that stand in a line across the PC board. The heatsink is not grounded and not supported in any other way. The speaker is mounted on a bracket behind perforations in the cover since the cover has to slide on when installed. The material that forms the front panel of the cabinet is soft aluminum, but it is secured at all four corners when the cover is in place. To increase the rigidity of the front panel, I used a piece of 0.062-inch (1.6-mm) aluminum cut to fit the front panel. It is held in place by the bushings of the various panel mounted components. It is also easier to paint and letter this flat plate and it covers four holes in the cabinet's panel intended for handles.

Because of the wide variations in tuning capacitors that might be used, mine wasn't mounted on the circuit board. Instead, it was mounted on the front panel of the cabinet and connected to the PC board with a short length of shielded wire. The details of the tuning mechanism are shown in **fig. 9**. This design evolved during my search for a simple slow-motion drive that is easy to duplicate. I salvaged the panel bushing and shaft from an open-style, rotary-wafer switch that had a standard 1/4-inch (6-mm) shaft. The detent mechanism and switch sections are removed, leaving only the bushing and shaft. The rear end of the shaft is then turned down to a diameter of 0.11 inch (2.8 mm). Notice the angled bevel where the shaft is turned down just behind the bushing. A 3-inch (76-mm) diameter disk is fashioned from 0.062-inch (1.6-mm) aluminum and secured to a collar which can be mounted on the tuning capacitor's shaft and held in place with a setscrew. Slide

COIL WINDING CHART

REF	BOTTOM VIEW	WINDING INFORMATION
T1		PRI - 16 TURNS NO. 40, TAP AT ONE TURN, 1.5 - 4.5μH SEC - 1 TURN NO. 40
T2		PRI - 16 TURNS NO. 40, TAP AT TWO TURNS, 1.6 - 4.6μH SEC - 2 TURNS NO. 40
T3, T9 T10, T12 T13		SEE TEXT
T4		SEE TEXT
T5, T8		PRI - 36 TURNS NO. 34, TAP AT 6 TURNS, 10 - 30μH SEC - 3 TURNS NO. 34
T6		PRI - 36 TURNS NO. 34, TAP AT 6 TURNS, 10 - 30μH SEC - 3 TURNS NO. 34
T7		PRI - 7 TURNS NO. 34 SEC - 1 TURN NO. 34 CORE - FERRITE BEAD AMIOON FB-43-101, FERROXCUBE K500100/3B, ETC.
T11		PRI - 16 TURNS NO. 40, TAP AT TWO TURNS, 1.5 - 4.5μH SEC - 2 TURNS NO. 40

REF	BOTTOM VIEW	WINDING INFORMATION
T14		PRI - 8 TURNS NO. 34, 0.4 - 1.2μH SEC - 2 TURNS, CENTER TAP
T15		PRI - 6 TURNS NO. 26, CENTER TAP SEC - 2 TURNS NO. 26, CENTER TAP FERRITE BALUN CORE
T16		PRI - 4 TURNS NO. 22, CENTER TAP SEC - 14 TURNS NO. 26 FERRITE BALUN CORE
L4		36 TURNS NO. 34, CENTER TAP, 10 - 30μH
L6		36 TURNS NO. 34, 10 - 30μH CUT OFF UNUSED PINS
L7, L8 L9, L10 L11		18 TURNS NO. 34, 3 - 7μH CUT OFF UNUSED PINS
L12, L13		3 TURNS NO. 26, CORE - 2 FERRITE BEADS END TO END AMIDON FB-43-101, FERROXCUBE K500100/3B, ETC.

the disk forward in contact with the beveled part of the shaft and tighten the setscrew. This creates a friction drive with a tuning rate of about 35 kHz per revolution and no backlash. The exact tuning rate depends on the distance from the shaft's center where contact is made with the disk. The hole for mounting the bushing is a little oversized, to allow for some adjustment. Frequency markings are applied to the disk and viewed through a window cut in the front panel. You can use a commercial vernier drive, if you have one, or as a last resort, use a small trimmer capacitor in parallel with a main tuning capacitor as a bandspread control.

component selection

There is little about the circuit that is critical. You can use substitutes for many of the components with little change in performance. I made an effort to minimize the number of component values used. Where possible, resistors and capacitors were selected from decade values in 1-3.3-10 sequence. The idea was to make it easier for those who had to buy what they didn't have. Most hams will resort to the junk box first. For resistors, watch ratios and keep the dc bias conditions for active devices from varying too much. For coupling and bypass capacitors, check the reactance at the frequency of interest and keep the leads especially short in rf circuits. If you use the circuit board layout provided, all resistors, except the bias resistor for the final, are quarter watt size. I used 1/4-inch (6.3-mm) disc ceramic and radial electrolytic capacitors throughout to minimize board size.

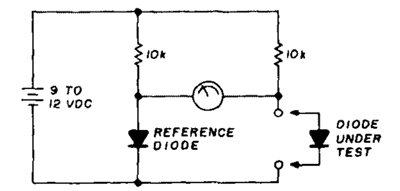


fig. 10. Use this circuit to match diodes *if you don't* have an accurate digital voltmeter. The reference diode is not to be one of the matched set unless you can reverse the positions of the diodes and still get a null.

You will have to wind a number of coils and transformers, the details of which are shown in the table. The variable inductances use 10-mm i-f cans that I rewound. They were originally 4.5-MHz and 10.7-MHz transformers in a selection from Poly-Paks. These higher frequency units have a plastic bobbin which makes them easy to rewind. This is not true of lower-frequency (455-kHz) transformers that use a threaded cup rather than a slug. Many of them have internal capacitors, and in some cases, I retained the internal capacitor and added enough capacitance externally to get near the value on the schematic.

Inexpensive, general-purpose transistors are used everywhere except in the final. The device that I used most was a 2N3904 (and its PNP complement, 2N3906) since it met my criteria of good performance, cost, and availability. It has the popular E-B-C lead configuration and there are a lot of substitutes that you can use without sacrificing performance,

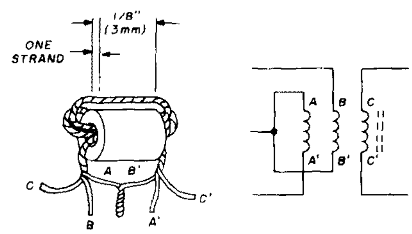


fig. 11. Typical mixer transformer. The wire should be twisted enough so you can count from 3 to 7 strands in a 1/8 inch (3 mm) length.

except possibly in the receiver front-end and transmitter driver. Remember to use transistors with low-output capacitance for Q1 and Q19. I was conservative with transistor ratings and used power-type transistors if there was a possibility of exceeding half the rating of a 2N3904.

The printed circuit layout provided accommodates the mini-dip package version of a 741 op-amp. The same layout can be used with the round metal can version if the leads are spread. Since a lot of op-amps are supplied with the same popular pinout as a 741, there are many substitutes that will work here, but make sure you use an internally compensated op-amp.

balanced mixers

The transceiver uses three double-balanced mixers. There's nothing hard about rolling your own. To get good carrier suppression, you simply have to maintain good balance. The two places where balance is important are the transformers and the diodes. Transformer balance is achieved by using transmission line techniques. Diode balance relies on matched diodes.

All of the diodes used in the transceiver, except one zener, are the same type, 1N4148 (1N914, 1N4454, etc.). You can get these diodes in quantities from mail-order houses for a nickel or less apiece.

Since you will need a total of twenty-seven for the rig, you'll have enough to find three matched sets of four. Because the highest frequency of interest is only 27 MHz, don't bother matching the reverse capacitance. Just use the same type from the same manufacturer, so the capacitance is close, and match the diodes' forward characteristic as outlined here. Measure and match the forward voltage drop across the diode with about 1 mA of bias current. A 10 K resistor in series with a stable 9 to 12 volt supply works fine. A high-impedance, accurate VTVM should be used (a digital meter is best), but don't despair if you don't have one. The bridge circuit shown in **fig. 10** works with any meter that can distinguish a change of a few microamperes.

The transformers for the transmitter and oscillator offset mixers (T9, T10, T12, T13) are the same. Using small enamel magnet wire, about AWG 34 (0.16mm), twist three strands together uniformly to make trifilar wire. Wind two turns of the trifilar wire on a ferrite bead (Ferroxcube K500100/3B, Ind. Gen. F1650-1-H, Amidon FB-43-101, etc.). Two of the strands are series connected to form the center-tap of the balanced winding (see **fig. 11**). The input transformer of the i-f mixer (T3) is constructed the same way except that two beads, stacked end-to-end, are used for the core. The output transformer of the i-f mixer (T4) is different because it matches the high impedance of the crystal filter. For this transformer, twist six strands of magnet wire together, and use two turns on a two-bead stack. Connect two strands to make the center-tapped winding, as before, then series connect the other four strands to make the high-impedance winding that connects to the crystal filter.

checkout and adjustment

Using these instructions and the theory of operation, you should have no trouble getting the rig up and running if you have had any experience with building. The first part of the circuit you want to check out is the 10-volt regulator, since this affects almost everything else. Connect the external 13.6-volt supply and monitor the input current, which should be about 100 mA for receive. The regulated output should be near 10 volts. Load the regulator's output and watch the voltage. It should change less than 10 mV for currents up to 750 mA, but go into current limiting at about 1 ampere. Adjust the value of R94 if the current limiting point is very far off.

Check the frequency response and peak gain of the CW audio bandpass filter using the adjustment points described in the theory of operation. A simple way to check out the entire audio circuit of the receiver is as follows: Set the bandwidth switch for **CW** and the **VOLUME** control at minimum, then connect a jumper from the speaker to the filter input (top of R43). When you turn up the volume, the whole loop will oscillate slightly (50-100 Hz) above the center frequency of the audio filter and the S-meter will track the volume setting.

The adjustment of the crystal filter and low-frequency crystal oscillator are interdependent. This is best done using a sweep generator. It's a slow, tedious job with an ordinary signal generator, but here's the procedure if a sweeper isn't available: Disable the AGC (lift one end of C45) and monitor the audio output, in the SSB mode, with an ac voltmeter. Start with L4 set to minimum inductance. Tune across the passband and you'll find nearly equal audio peaks at 500 and 2000 Hz with a big dip in between. Adjust the tuning of the 3.58 MHz crystal oscillator if necessary to get the frequencies near these values. Increase the inductance of L4 until the variation of audio across the passband is less than about 1 dB from 350 to 2200 Hz and there is a sharp roll-off of the audio at 300 Hz. Don't confuse the low-frequency roll-off of the audio circuits for the filter's corner. To be sure, you can tune the crystal oscillator up a little higher than normal, which will shift the audio frequencies up, so that the filter's corner frequency can be identified, then reset C54 so the corner is at 300 Hz. Set the bandwidth switch to **CW** and center the signal at the peak of the audio filter. Temporarily short the collector of Q19 to ground and adjust C53 for a zero-beat by watching the voltage at the collector of Q11 with a scope.

If you have trouble getting the 3.58-MHz oscillator adjusted properly, the problem is probably a transistor with too much junction capacitance. One possible solution is to bias the junction of Q19 by connecting a high-value resistor (10 K to 100 K) from the collector to +10 volts. You can check the operating frequency of each crystal in the circuit and use the one with the highest frequency in the oscillator.

The other adjustments consist of tuning for maximum output at the center of the band.

On-the-air reports with the rig running barefoot are favorable. A number of operators have commented on the good sounding audio quality that comes through in spite of QRP signal levels. I get questions about what kind of processor I'm using. The passband may be only 2 kHz, but only once did I get a comment about the lack of highs in the signal. My voice tends to be on the low side anyway. I've had no trouble working stations all over North and Central America from my home in Tennessee. Pile-ups on DX stations are another matter however. Of course, a linear amplifier is next on the want list.

ham radio

Bobtail curtain follow-up: practical DX signal gain

The second part of a two-part series on this remarkable antenna

The actual DX signal gain of any one type of antenna over another, at distances beyond about 2500 miles, does not always correlate well with the theoretical "free space gain over isotropic." After all, antennas do not operate in free space. Surrounding objects, especially ground, are a part of the antenna system.

For distances beyond 2500 miles, angles of signal *departure* below about 15 degrees are almost always the most effective. This is true regardless of propagation path, whether it's one acute geometric bend near midpoint or a chordal or ducting mode. And it's true regardless of ionospheric tilt. Although it's most noticeable on 10 and 15 meters, it still applies to long-haul 40- and 75/80-meter *propagation*.

To get the angle of radiation down while still keeping the antenna height acceptable on 40 and 75/80 meters, vertical antennas have long been used. Some verticals do a good job. A few, such as a full-size half-wave vertical, can do an excellent job in all directions. Others seem to radiate "equally poorly in all directions." I'll not take time to go into all the reasons why short vertical radiators that are current fed near ground level are often ineffective.

As noted in Part I of this article, simply turning the antenna upside-down greatly reduces the ground loss problems. For one thing, it minimizes the conduction current flowing to ground at the feed point when the antenna is ground mounted. It also minimizes the losses caused by displacement currents in the near field fighting their way through lossy dirt trying to find a "mirror image" that, in this case, is more theoretical than real. In addition, getting the high-current portion up in the air allows the antenna to radiate somewhat more effectively, and it lowers the angle of radiation slightly.

Users of the Bobtail antenna often report gain improvements on long-haul DX of from 10 to 20 dB over their previous antenna. But only a little of that improvement results from the azimuthal directivity. The sometimes startling effectiveness of the Bobtail for 40- and 75/80-meter DX is the result of its inverted configuration. This is ordinarily more noticeable in a built-up, residential area than in an open field.

The gain attributable to the horizontal directivity of a two-element Bobtail, or half square, is about 4 dB over that of an inverted groundplane using two resonant radials. The half-power beamwidth of each lobe of the bi-directional figure-8 pattern is about 60 de-

By Woodrow Smith, W6BCX, 2117 Elden Avenue, Apt. 20, Costa Mesa, California 92627

grees — wide enough to cover some worthwhile geography while still providing useful gain.

The full-size Bobtail (for those that have the room) has a directivity gain of slightly over 5 dB (compared with the same reference). The half-power beamwidth is about 50 degrees. Four or five dB doesn't sound like much, but if your signal is marginal it can make the difference between copy and no copy. On receive, the discrimination you get from the azimuthal directivity can be worth more than the 4 or 5 dB when it comes to what you can copy through noise and QRM.

The slight extra gain of the full-size Bobtail over the two-element should be considered a bonus. The main advantage of the full-size version is that spurious lobes are reduced in amplitude, and therefore end-fire, high-angle pickup on receive is reduced. There is no point in being able to deliver a readable signal 6000 miles away on 40 if you can't hear the other station because of bad off-axis QRM from a station 600 miles away.

direct coax feed versus voltage feed

As noted in Part I, if certain precautions are taken, direct coax feed can be used with a half-square version of the Bobtail. The disadvantage of doing so is that it limits the antenna to one band. Voltage feeding one end of a half square permits use on half frequency, as sort of a drooping, half-wave dipole. At twice the frequency it functions as a combination of two co-phased vertical dipoles a wavelength apart, with the resultant cloverleaf end-fire and broadside pattern having its nulls filled in fairly well by the pattern produced by the horizontal full-wave portion. Voltage feeding the half square at three times frequency produces an interesting multiple lobe pattern that results in unpredictable results. Voltage feeding one end of a half square cut for 40 meters will thus provide a low-angle, "far DX" figure-8 pattern on 40 while also doing a good general-coverage short- and medium-haul job on 80.

On 20 meters the composite pattern obtained results in a good, general-coverage "long" DX antenna that in addition is effective in the 750- to 1500-mile range. Voltage feeding it on 15 meters produces an interesting multiple lobe pattern that will sometimes, as noted above, produce results that often are surprising — and occasionally amazing.

On 10 meters it should be considered simply a random long wire that is capable of providing lots of good clean fun when the band is hot.

Unfortunately there is no really good way to feed a three-element Bobtail directly with coax, even for one-band operation. No matter how the coax is brought down from the center element there is going to be objectionable unbalanced coupling to the coax from radiating portions of the antenna. Using a balun does not cure the problem.

Connecting the coax to an end radiator junction as described for the half square does reduce the unbalanced coupling, but the current distribution in the three elements no longer is symmetrical. Current will be greatest in the fed element, thereby skewing the pattern and reducing the gain.

"Here's the best place to feed a Bobtail curtain . . ."

The only really good way to feed a three tailer is to voltage feed the bottom of the center element. This permits multi-band operation in pretty much the same fashion as with the half square, though the lobe pattern on bands other than the fundamental will be slightly different. The main difference is that on half frequency the three-tail version makes a much better end-fire "medium DX" antenna than a half square. However, this is at the expense of high angle (short haul) effectiveness, particularly broadside.

"Zepp" voltage feed

Way back in the 1930s, PA0ZN came up with an antenna resembling a two-tail Bobtail (or half square) in appearance except for feed method. It was fed at the center of the horizontal section (a high impedance point) via Zepp feeders. The antenna obviously belongs to the "inverted ground plane" family, with the attendant advantages over a right-side-up arrangement.

However, the mutual impedance between the Zepp feed line and the rest of the antenna is such that strong in-phase currents are induced in the line, thus producing considerable "antenna effect" on the feed line. This is strong enough to produce sufficient distortion of the pattern to "dirty up the nulls" a bit without providing a significant increase in gain. The pattern is cleaner when the Zepp feed line is attached to one end of the radiator. While both radiator and

feed line are unbalanced slightly with such an asymmetrical arrangement, the resulting imbalance is not enough to cause serious mischief.

If you want a dual band (say, 40 and 20 meters) omnidirectional DX antenna and would like to use Zepp feeders all the way to the shack, I would suggest an inverted ground plane with two radials, with the vertical element of the I.G.P. fed at the bottom with one side of the open wire line.

If you would like to Zepp feed a three-element Bobtail, be sure to feed the center element. But remember that the main reason for using a three-element Bobtail instead of a half square is the cleaner pattern, and while Zepp feeding the center element will not unbalance the antenna itself, the inherent unbalance in the Zepp feed line will cause some pick-up by the line when receiving, even if the open line is brought off at right angles out of the near field.

Considerable voltage will be built up across an open feed line at the high impedance points when the line is used as a Zepp feed line on transmit, so be sure to use sufficient spacing if you are running power.

ground screens and grounding

No antenna that is fed via an unsymmetrical feed system can be 100 percent ground-independent. However, if the impedance between antenna feed point and ground is over about 1000 ohms there will be very little current flowing to ground. In this case not much of an earth ground will be needed at that point for the antenna. A small ground screen, laid on the ground or a flat roof, or suspended or supported near the network matching 50-ohm coaxial line to the high impedance Bobtail antenna feed point, makes an effective rf ground. Such a screen will make a better rf ground than a lightning stake, or stakes driven in the soil. If one of the latter, properly installed, is advisable for lightning protection in your area, it still is a good idea to use a small ground screen as an rf ground.

One ready-made ground screen widely available is a 3 × 5 foot (0.9 × 1.5 m) piece of galvanized hardware cloth, packaged by Sears under the catalog number 44531 and available off-the-shelf at some Sears retail stores. Current catalog price is about $7.00 a roll.

To see if the ground screen is doing its job, simply touch the "rf ground" while running low power to see if the VSWR or field strength changes significantly. If it does, either shorten the connecting wire between screen and matching network or series resonate the wire with a mica capacitor. The value will not be especially critical. Possibly on 75/80 more screen will be required. Ordinarily if the antenna dimensions are near optimum, not much screen will be required in order to do the job properly.

10-MHz Bobtail dimensions

Because adding 10 MHz to a typical tribander is not the easiest thing in the world to accomplish, there is bound to be interest in 10-MHz Bobtails and half squares. Dimensions are not extremely critical. Except for direct coax feed of a half square, slight deviations from optimum can be compensated for in the tuner or matcher at the antenna end of the coax line.

Assuming No. 12 or 14 wire (M2.1 or 1.6) and typical insulators, the following dimensions will be found satisfactory for 10.1 to 10.15 MHz:

spacing of vertical elements:	48 feet 9 inches (14.86 m)
length of vertical elements:	23 feet 7 inches (7.19 m)

For the sake of convenience the voltage fed element can be made up to 5 percent shorter or 8 percent longer if fed by a resonant tank or L network. For Zepp feed, the fed element should be cut to the exact length shown.

guys and metal masts

When using wooden masts to support a Bobtail the usual precautions apply to breaking up adjacent guy wires to avoid resonance. There is nothing wrong with using metal masts of EMT or aluminum tubing as the outside vertical elements to save lot space. This requires a strong base insulator of low loss material, preferably having low shunt capacity. Nonmetallic guys do a better job electrically than guy wires, even if the latter are well broken up. The tubing should be of no greater diameter than what is required for adequate mechanical strength. A tubing element will be slightly shorter physically than a wire element for the same electrical length. Also, the shunt capacity of the base insulator will require shortening the element a bit more. However, if you merely make them 3 percent shorter than a wire element of optimum electrical length, and try to keep the base shunt capacity low, it should be close enough. Tubing joints should make good electrical contact or be jumpered.

Unless a piece of thick-walled fiberglass tubing or the like is used as the base insulator, it should be used only to support the weight of the mast, and not to hold the bottom rigidly vertical. This probably means an extra set of guys.

Keep in mind when planning your installation that the bottoms of the tails are quite hot with rf and can cause a bad burn if you are running much power. One way to avoid this is to slip some small-diameter clear plastic (such as fuel line) tubing over the ends of wire tails closer than 7 feet (2.5 m) from ground. If you decide to use metal tubing for the end elements, slit or saw lengthwise a few feet of PVC plastic pipe and tape it over the bottom of each mast.

ham radio

A simple, automatic negative voltage tracking system

dual voltage power supply

Anyone who tinkers with integrated-circuit projects will eventually have to provide a negative voltage supply for op amps (operational amplifiers — see **fig. 1**). Batteries may seem a simple solution, however, after purchasing several quality 9-volt alkaline batteries at $2 each, you soon see the false economy *in* that.

Deciding my money could be better spent on parts for a power supply, I set out to find a good schematic design. After my initial search it seemed I would have to purchase two voltage regulators, one for positive and one for negative, to obtain a regulated positive/negative supply. I wasn't willing to spend at least $4 each for voltage regulators and have the hassle of adjusting two knobs to set both voltages. Neither was I excited about watching two meters to be sure both voltages remained equal under changing load conditions. After extensive research, I found a small diagram in a Radio Shack semiconductor guide under the 78XX series voltage regulators using op amps.

A 741 op amp is much cheaper than a monolithic negative voltage regulator, such as the LM337T. A 741 op amp costs 79¢ at Radio Shack (part #276-007). Even if you add the cost of Q3 to the op amp, it is still three dollars cheaper than the negative regulator, plus the 5-kilohm potentiometer required to control it. And, only a single control is needed to set both voltages simultaneously.

operation

Automatic negative voltage regulation for a fixed-voltage supply is provided by the circuit illustrated by **fig. 2**. It uses a 741 op amp in an inverting configuration. The op amp compares the voltage between pins 2 and 3. It supplies an equal but opposite voltage at its output, pin 6, by sensing the difference between the two input voltages and inverting that voltage at its output. This in turn biases the pass transistor into conduction providing the power supply negative voltage output.

Fig. 3A shows how this basic idea is used to build a positive/negative variable voltage supply. R11 and R12 are current-limiting resistors connecting the negative and positive power supply outputs. Since these resistors are of equal value, they will have equal voltage drops with equal applied voltages.

When the negative supply voltage is exactly equal and opposite the positive supply voltage (− 10 volts and + 10 volts, respectively) the two voltage drops across R11 and R12 are equal. They cancel each other and provide zero volts at pin 2 of the op amp, U1, which compares the zero volts on pin 2 with the ground on pin 3 (also zero volts) and finds they are equal. U1 therefore gives no output on pin 6.

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The finished product. The top half of the faceplate contains all controls for the 1-ampere supply. The bottom half has all controls for the 2.5-ampere supply. The face plate was made prior to final testing, and bears the estimated 3-ampere rating.

When adjusting the positive voltage regulator (U2) for a desired positive supply voltage, that voltage drops across R11, raising the voltage sensed by pin 2 of U1. U1 supplies an equal but opposite voltage to the base of Q3, biasing Q3 into conduction. Q3 drives Q2 into conduction, providing a negative voltage at the supply output. There it drops across R12, lowering the voltage sensed by pin 2. When the negative supply voltage lowers to the exact opposite of the positive supply voltage, they cancel each other at pin 2, completing the cycle. The op amp continuously and instantaneously controls the negative supply voltage, keeping it exactly equal but opposite the positive supply voltage.

circuit description

R13 is a balance control to adjust lor differences in the values of R11 and R12. A 1 kilohm potentiometer should be sufficient for R13, since the difference between two 4.7K, 10 percent tolerance resistors shouldn't exceed 940 ohms, even in an extreme case. I used a circuit-board-mounted pot since this should be a one-time only adjustment (when you first calibrate the supply). Simply adjust R13 so the negative voltage is exactly equal to (and of opposite sign to) the positive regulated voltage. No further adjustment should be required. I suggest calibrating R13 in the VOM range, that covers the full power supply voltage range. This will eliminate the need for switching ranges when checking your calibration. Switching from one range to another could give you slightly different readings at the same voltage setting.

R10 is a current-limiting resistor intended to keep U1 from burning itself out should it try to supply too much current to Q3. When testing the circuit, I tried to bias Q2 directly with U1. But Q2 required too much base current and U1 self-destructed. Any small current PNP can be used in the place of the Radio Shack 2025 I used for Q3, as long as it will withstand the required driving current to Q2. I measured 19 μA on the base of Q2, under *no load*, and 17 mA (nearly 900 times more current) at a 2.4 ampere load at 15 volts output.

The op amp requires a positive/negative voltage supply for its own operation. The regulated supply voltages to run the op amp are obtained by employing 18-volt zener diodes D1 and D2. Since the op amp draws very little current, R3 and R4 are mainly used to limit current through the zeners. Half-watt or one-watt zeners provide a sufficient power rating.

a new approach

I believe the manner in which I have employed the regulator U2 in biasing the pass transistor Q1 is a novel approach. Other designs using 3-pin adjustable or 3-pin fixed voltage 78XX series regulators use a PNP pass transistor with its base connected to the emitter through a resistor or its base tied to the regulator input, as in fig. 4. The regulator output and pass transistor collector are tied together at the output. My approach uses an LM317T 3-pin adjustable positive voltage regulator (U2). It can be used alone for up to 1.5 amperes current, or with a pass transistor to increase the available current. I used this type regulator alone for the 1.2 to 25-volt supply, as shown in fig. 3B. This extra supply was built into the same cabinet as the positive/negative supply. U3 is shown in a standard application of the LM317-type regulator. I used 0.47 μF electrolytics for C5, C6, C9, and C10. The recommended capacitors are 0.33 μF tantalums for C5 and C9, and 0.1 μF tantalums for C6 and C10. (See final comments.)

In fig. 3A you see that the output of U2 is used to bias the base of the NPN pass transistor directly, as opposed to the bias coming from the same line as the PNP emitter (fig. 4). R5 and R6 are used as current dividers. R5 is half the ohmic value of R6, allowing twice as much current through pass transistor Q1 as passes through regulator U2. Therefore, if the regulator is passing its maximum current of 1.5 amperes, the transistor can pass twice that, or 3 amperes, the-

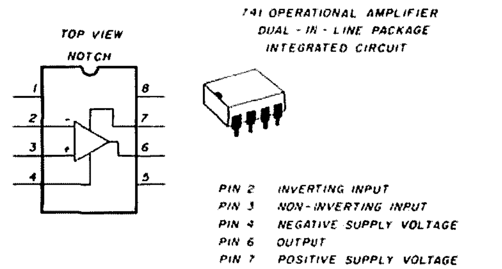


fig. 1. 741 op amp dual in-line package integrated circuit.

oretically providing up to 4.5 amperes of current at the output.

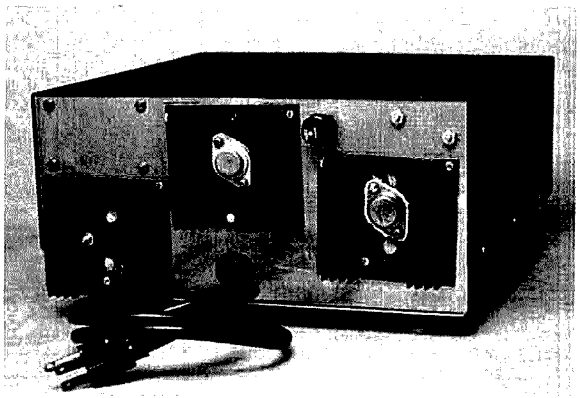
The LM317-type regulator has internally controlled automatic shutdown for overheating and overcurrent protection. Should the regulator reach its safe limit, it will automatically shut down. In this configuration, if U2 shuts down, there is no bias for Q1. This shuts off Q1, and consequently, turns off the negative supply. Q1 is a 115-watt transistor rated at 15 amperes collector current, so it is not likely to overheat while passing only 3 amperes of current. In fact, it should never get past warm to the touch if it is properly heatsinked. Even if the regulator is operating at its extreme limitations, it will always shut down in an overload situation before Q1 can overheat, because Q1 is operating at only one-fifth its rated capacity in that same extreme case. Overdesign was mostly a function of junk box availability.

D4 is used to isolate the Q1 emitter from its base. If D4 shorts out, it shuts down Q1, again preventing any possibility of thermal runaway. Be sure D4 is rated higher than your expected 1.5 ampere current through the regulator, or you may be constantly replacing D4. A 2.5 ampere diode should provide a sufficient safety margin.

As shown in **figs. 3A** and **3B**, I used fused outputs on both power supplies as an additional safety factor.

additional features

M1 and M2 are two halves of a dual VU meter modified for use as voltmeters. R14 and R15 are actually 100-kilohm resistors in series with 25-kilohm circuit-board-type potentiometers, used to calibrate the meters. I added some LED indicators (D3, D6, and D7). D6 and D7 indicate (**fig. 3B**) to which voltage scale I have the meter set. I use a 0-10-volt scale for better resolution at low voltages, and a 0-30-volt



This rear view shows heatsinking. Far left heatsink is for the 1-ampere supply regulator, mounted to the sink from the inside. Center sink has the GE 14 (Q1) while the right sink has the 2SB539 (Q2). Be sure to use silicone heatsink compound when mounting transistors to any heatsink.

scale for higher voltages. A microammeter would be preferred over the milliammeter used for M3, but again, I used what was available in my junk box. R20 through R23 are calibrating resistors for M3.

D3 tells me when the output is turned on by switch 2. Since the LM317 regulator can be turned down to no lower than 1.2 volts, such a switch is handy. With S2 open, U2 has no input and, therefore, no output. Likewise, Q1 and Q2 are turned off when U2 has no output, so there is no need for a switch on the negative side.

My design has partially taken care of the minimum voltage problem, even without S2. Due to the voltage drop across D4, and a similar drop across the Q1 base, the minimum output is reduced from 1.2 to about 0.6 to 0.8 volts. That low a voltage forces very little current through anything but a dead short. However, you can program the minimum voltage up to 1.2 volts, or any higher voltage, by adding a fixed resistor between R8 (or R18) and ground. The formula for calculating the needed resistance is covered in the applications data sheet supplied with the regulators.

a limitation

The idea of using an op amp as the negative voltage regulator has one limitation: the supply voltages for the op amp. The operating voltage limits for op amps are generally from 16 to 18 volts. The op amp has internal voltage drops. With the extra voltage drops across R10, Q3, and Q2, I can get only up to 15.5 volts at the negative supply output (depending on the load). This is quite sufficient for operating op amp test circuits.

The 741 op amp I used was rated at 16 volts maximum supply voltage and I am presently operating at

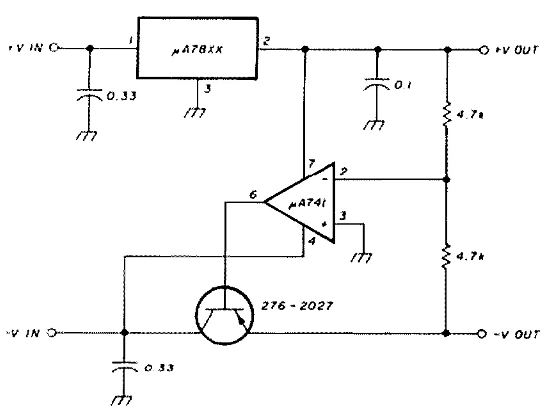


fig. 2. Automatic negative voltage regulation circuit, copyright Tandy Corporation, *1980 Archer Semiconductor Guide*, reproduced with permission.

18 volts. No damage has occurred thus far, but I would not be surprised if the op amp fails one day. Since this is a learning experience for me, I'm willing to chance a 79-cent op amp. But if you want to be safe, use the recommended supply voltages when choosing your zeners.

performance

My original intent was to operate small experimental circuits which draw no more than 500 mA in extreme cases. The power supplies outlined here are more than adequate for that purpose. I built them to

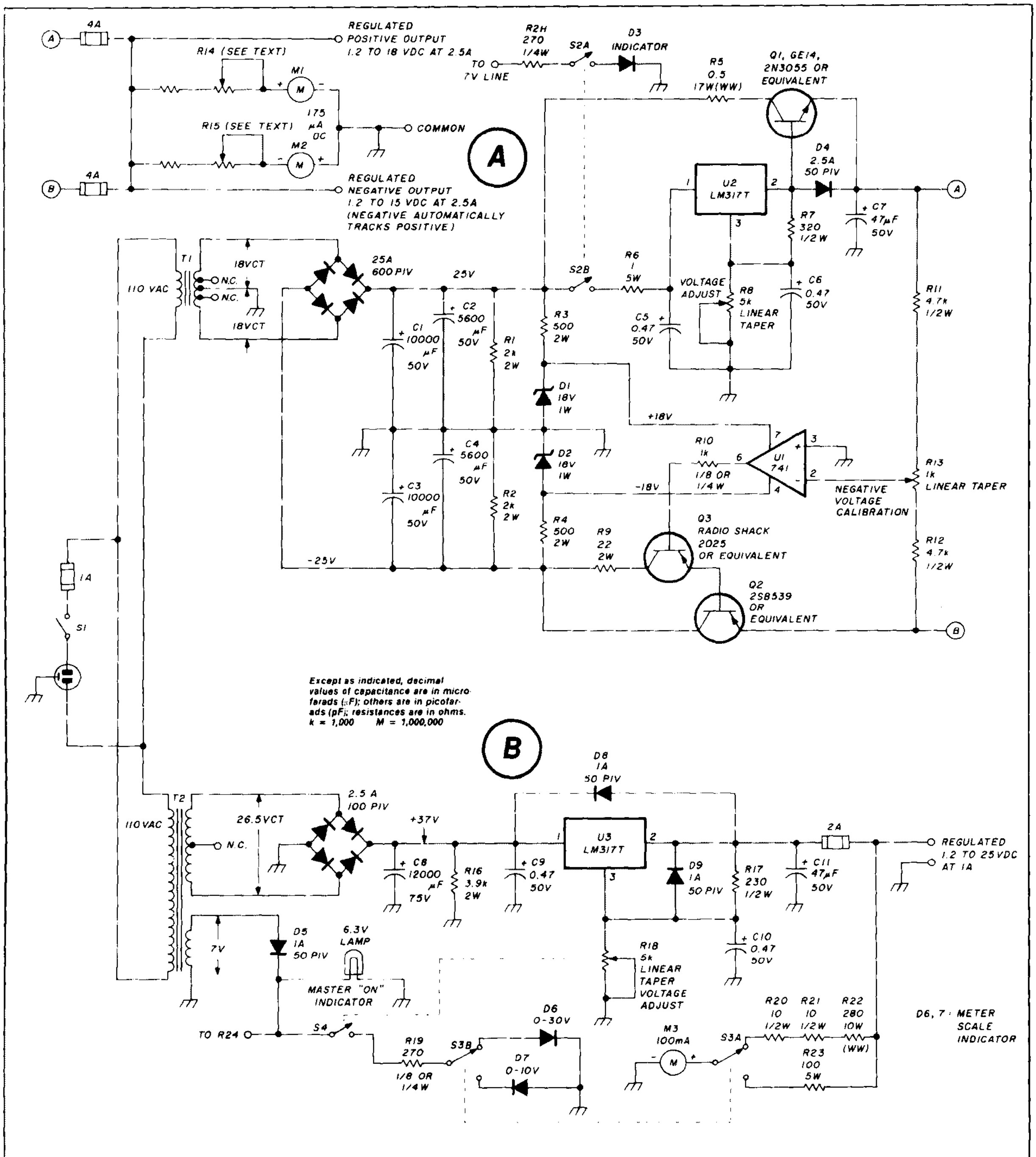


fig. 3. (A) Positive/negative variable voltage supply. (B) An LM317T three-pin adjustable positive voltage regulator is used in the 1.2 to 25-volt supply.

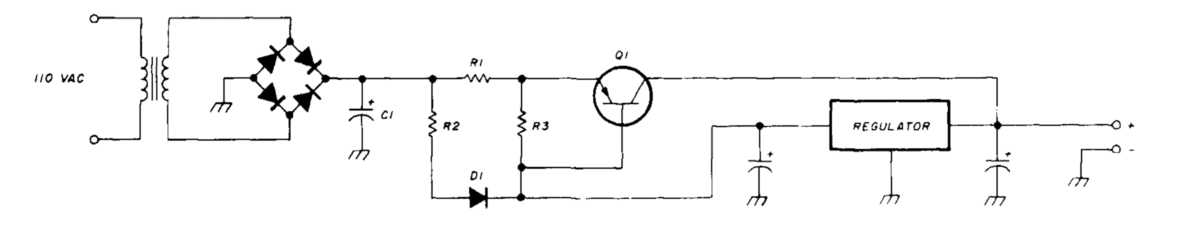


fig. 4. This is the basic circuit popularly used for providing extended current capability with a 3-pin voltage regulator. R3 is used in some design variations.

give everything the transformers would provide, and they do.

Voltage and ohmic measurements were made using a Fluke 8024A Digital Multimeter. Ripple measurements were made with a Hewlett Packard 180A 50-MHz scope, with the Hewlett Packard 1801A Dual Channel Vertical Amp and 1821A Timebase.

I overestimated the capacity of T1, a transformer I salvaged from an old TV set. I guessed it would supply 3 amperes, which was about 0.5 ampere too much. T2 was salvaged from a junked receiver and rewound to suit my needs.

The positive supply in **fig. 3A** showed 17.4 volts into a 6.3-ohm load, with less than 10-mV ripple. However, between 17.4 and 17.8-volts output the ripple rose sharply to 900 mV. That gives 2.76 and 2.82 amperes at 17.4 and 17.8-volts respectively. So, a rating of 2.5 amperes for T1 would have been quite reasonable, instead of my estimated 3 amperes.

The negative voltage side of **fig. 3A** delivered 14.6 volts into the same load with less than 10-mV ripple also. Since the negative voltage is limited by the op amp supply voltages, it would not adjust any higher.

Positive and negative outputs were tested simultaneously into 12-ohm loads, and yielded the same performance as their individual ratings into 6.3 ohms. Additionally, the negative voltage tracked positive within ±0.05 volt throughout the range 1.2 to 15 volts.

The positive supply was further tested using a 3-ohm load, which put a 4.5-ampere current through the supply. The regulator circuit was monitored with one meter while the total output was monitored on another. The regulator circuit provided 1.6 amperes of the total 4.5-ampere output, with 2.9 amperes flowing through the pass transistor. This is very close to the designed current distribution ratio. The supply was run for a full fifteen minutes at the 4.5-ampere output, which is 2 amperes higher than its rated capacity. My intent was to force the regulator to overheat and shut down. However, the supply performed flawlessly, and ran the whole time with no signs of cutting back or shutting down. The regulator heatsink was too hot to touch but the pass transistor was not excessively hot. No doubt forcing the issue would eventually cause the regulator to shut down. Since the unit had already performed far above expectation, I decided to let it go at that.

The supply in **fig. 3B** works equally well. It supplies 24 volts into an 18.3-ohm load with less than 10-mV ripple. At 24 and 25-volts output, with the same load, the ripple sharply rises to 900 mV. The 1 ampere rating for T2 is therefore a fair rating. This supply was also tested with a 31-ohm load and yielded 27.2 volts at less than 10-mV ripple. This is surprising since that is above the transformer voltage of 26.5 volts.

final comments

A technician friend reviewed my test data and recommended changes in C5, C6, C9, and C10, which should improve the ripple rejection. Due to the internal reactances of electrolytic capacitors, as compared to tantalum capacitors, the desired frequency response can be approximated by using 0.1-μF ceramics for C5 and C9, and by using 10-μF electrolytics for C6 and C10. You might try this change when building yours.

Except for the LM317 regulators, all parts for these two power supplies were obtained from the proverbial junk box. I present these ideas as building blocks for others to improve upon. I am certainly no expert, but I will try to answer any questions you may have if you will enclose an SASE. I suggest some type of overvoltage protection circuit if you plan to use this or any other power supply with expensive or sensitive equipment.[1]

I wish to thank N4BGU for assisting with final testing and test data, and without whose patience I would probably not have learned much of anything about electronics except that required to pass my Novice exam.

reference

1. Evert Fruitman, W7RXV, "A Better Overvoltage Protection Circuit," *73*, January, 1980, page 140.

ham radio

ham radio magazine presents this latest examination of GaAs FET low-noise operation at *lower* than normal frequencies.

GaAs FET performance evaluation

and preamplifier application

A look at some popular devices for UHF and VHF

The performance of GaAs FETs in terms of noise figure, gain, and overload capabilities at VHF and above is well known. A new device, the ALF1023 by Alpha Industries, makes it possible to obtain good performance at low cost ($12 in small quantities). These new GaAs FETs are particularly useful in improving two-meter receivers where front-end overload is a problem.

I tested seven popular GaAs FET devices at 144 MHz for noise figure, gain, 1-dB compression point, and third-order output intercepts. The results are shown in **table 1**. All the devices have essentially the same characteristics at 144 MHz.

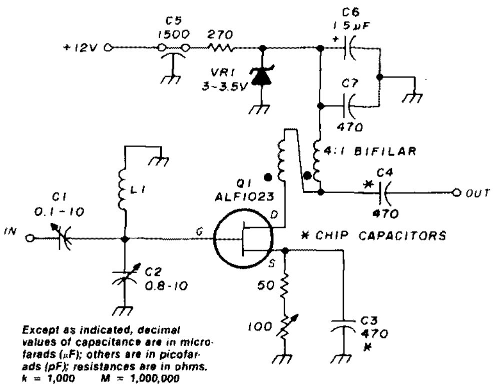


fig. 1. Preamplifier circuit for the ALF1023 GaAs FET. Values shown are for 144 MHz.

The author, Dennis Mitchell, K8UR, is employed as Applications Engineer for Alpha Industries.

The third-order output-intercept point was measured under slightly different bias conditions than for low-noise biasing, as was the 1-dB compression-point measurement. A higher drain current (approximately 50 percent higher) was used for these measurements. The third-order output intercept was substantially improved by this bias condition.

This improvement in third-order intercept indicates one of the design trade-offs that one must consider. A sacrifice of noise figure to obtain a higher intercept point is typical. However, when biased for lowest noise, the GaAs FET is already superior to the bipolar device it is to replace.

The improvement in freedom from overload and intermod obtained by using a GaAs FET device to replace a bipolar transistor is pronounced. **Fig. 1** shows the schematic of the test circuit used at 144 MHz.

Similar circuits for 220 and 432 MHz are popular and give similar results. At 432 MHz and above, however, some devices start to show slightly higher noise figures than others, even though they are essentially identical at 144 MHz.

The accurcy of noise-figure measurement must be mentioned here, for I have encountered advertisements stating "0.2 dB" noise figure. Noise-figure measurements are relatively easy to make — noise-figure accuracies are inherently uncertain. This uncertainty will not be explored here except to state that the major elements of this error are:

1. Instrumentation error (meter, electronics accuracy)
2. ENR uncertainty (noise-source accuracies)

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3. Mismatch uncertainties (VSWR between source and device)

4. Second-stage uncertainties

I did not verify the noise-figure claims for many Amateur-band preamplifiers tested.

The system used for the gain and noise-figure measurements listed in **table 1** was an HP 346B Noise source and HP 8970A Noise-Figure Meter with a typical root-sum-of-squares uncertainty of ±0.23 dB at 144 MHz.

Fig. 2 shows a plot of the S_{11} and S_{22} characteristics of the ALF1023 GaAs FET from 100 to 8000 MHz. **Fig. 3** clearly depicts the ALF1023 dc and rf performance. **Fig. 3a** is a curve trace showing I_{ds} versus V_{gs} with an I_{dss} of 50 mA. **Fig. 3b** illustrates the trade-off possible between noise figure and intercept point for given values of drain current.

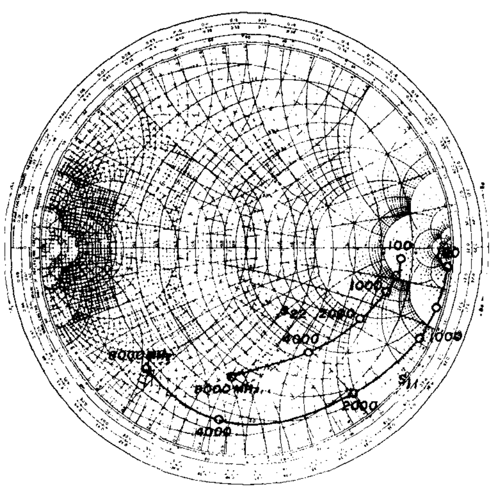
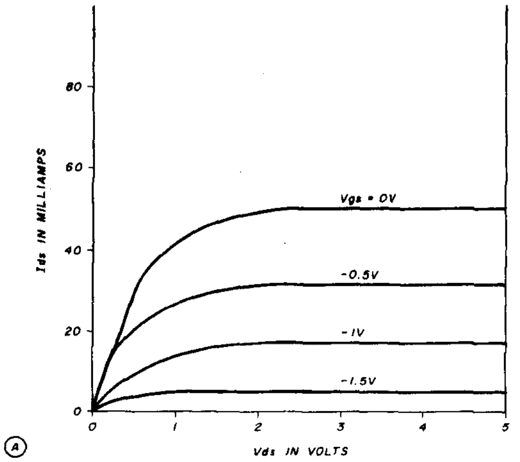


fig. 2. Plots of the S_{11} and S_{22} characteristics of the ALF1023 GaAs FET.

table 1. A listing of test results of some popular devices for VHF and UHF preamplifiers, including the new Alpha Industries ALF1023.

device	noise figure dB	gain dB	1-dB compression point, dBm	3rd-order output, intercept, dBm	drain bias	approximate (1-100) cost, $
Avantek AT-8110	0.6	19.0	+18	+28	3 V, 20 mA	44
NEC NE21889	0.6	19.9	+17	+27	3 V, 12 mA	30
NEC NE72089	0.6	19.8	+17	+27	3 V, 12 mA	12
MGF 1402	0.6	19.8	+17	+29	3 V, 24 mA	30
MGF 1400	0.6	19.7	+17	+29	3 V, 25 mA	22
MGF 1200	0.8	19.8	+15	+30	3 V, 25 mA	15
ALF 1023	0.6	19.8	+19	+30	3.5 V, 15 mA	12



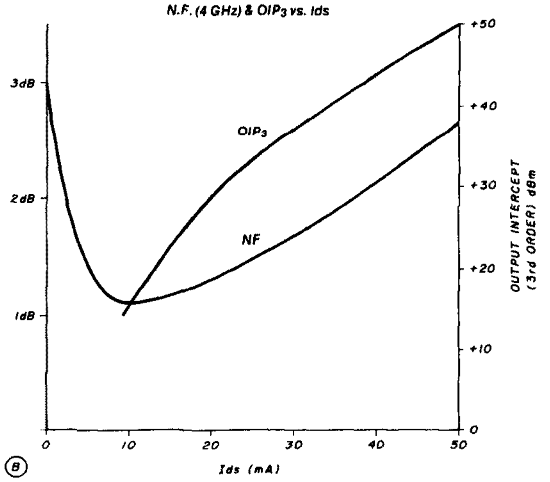


fig. 3A. DC curve trace showing drain current versus gate voltage for the ALF1023. fig. 3B. Tradeoff between noise figure and intercept point versus drain current.

technical forum

Welcome to the *ham radio* Technical Forum. The purpose of this feature is to help you, the reader, find answers to your questions, and to give you a chance to answer the questions of your fellow Radio Amateurs. Do you have a question? Send it in!

In this month's Technical Forum, we are going to begin presenting questions for our readers to answer. Our readers are encouraged both to send in questions of their own relating to Amateur Radio, and to answer the questions of other Amateurs as they appear in this column. The best and most complete answer to each question will be published in a future issue, and the reader who provides the answer will be awarded a prize in the form of a book from Ham Radio's Bookstore. Send all Technical Forum correspondence to Technical Forum, Ham Radio Magazine, Greenville, NH 03048. Now for the first question:

I am building an eight-turn helical antenna for the 432-MHz band. Every article I have read indicates a 140-ohm input impedance (or something close to that value). My feed line is TV coax cable: 70-ohm solid aluminum outer jacket with solid, copper-coated steel wire for the center conductor and foam insulation.

Can anyone tell me the most efficient way to couple the antenna to the 70-ohm line? And can anyone tell me the best way to couple the 70-ohm line to the 50-ohm input of the convertor?

Joseph Czerniak, W8NWU

At the Dominion Radio Astrophysical Observatory, British Columbia, Canada, we have built a synthesis radio telescope, using surplus 28- and 30-foot reflectors (see below). Aperture synthesis is a technique for making big radio telescopes out of little ones, and we have four small dishes spaced along a 2000-foot baseline. This gives our synthesis telescope a beam of 1 minute of arc (one sixtieth of a degree) at 21 cm wavelength (1420 MHz). It is a powerful radio telescope, which produces world-class astronomical research.

The four dishes we are using were obtained from surplus. Two 28-foot Kennedy Model 717 antennas came from the Canadian Defense Research Board. Two 30-foot Radiation Systems dishes were found in a scrap metal yard in California.

We are looking for more dishes of these two types, so that we can extend our telescope and make it more sensitive. Can any of ***ham radio****'s readers help us locate used dishes which might be for sale?*

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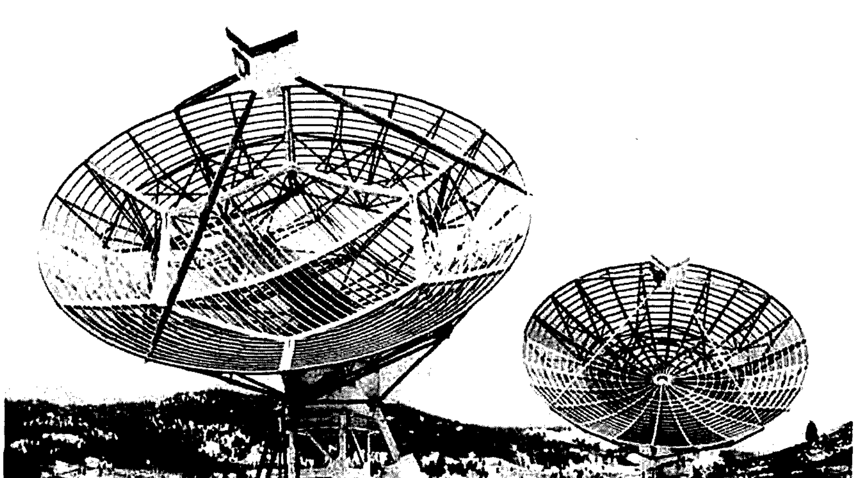


ham radio TECHNIQUES

Bill Orr W6SAI

RFI revisited

In my September, 1982, column I discussed some of the problems I encountered with severe radio and TV interference generated by my microwave oven. I also told of the subsequent runaround I received from the dealer and the manufacturer's service department. I've received a lot of mail about microwave oven RFI, and I think you might be interested in some of the remarks:

Art Nichols, W6EVL — "I returned my first Montgomery Ward microwave oven because it was 40 dB over S9 on 80 meters. The second one I got was better — it was only 20 dB over. I used an external filter until the warranty period was over, then put a pair of 0.1-μF, 600-volt capacitors from each side of the line to the case. That did the job fine. Montgomery Ward told me they didn't supply a filter."

David Cornell, K9BO — "I live in a medium-signal-strength area outside of St. Louis, Missouri, and own a General Electric Model J-ET015-OY1 oven. The TV shows no interference from the oven, and all I hear is a slight noise on the broadcast band when the radio's tuned to a quiet spot. This contrasts with your Amana report, suggesting that perhaps some microwave ovens are less objectionable than others in terms of RFI."

"He finally wrote the FCC about the problem, and received a reply by telephone."

Jim Ford, N6JF — "I was more successful than you in getting a filter for my microwave oven, but it took a letter to the Bureau of Appliance Repairs in Sacramento, California. The filter they installed helped but was no cure. A friend's Sharp microwave oven is clean of RFI. There is no interference to any channel of a nearby TV set that uses a set of rabbit-ears for an antenna." (Note: Jim's oven is an Amana. The form letters he received from the Customer Specialist at Amana seemed to indicate that they either don't understand the problem or else prefer to pretend it doesn't exist.)

Francis Stifter, President, Electronic Specialists, Inc. — "We have had many calls from people with microwave oven interference problems. One of our line-cord filters usually takes care of it. Occasionally it is necessary to fold the oven power cord into a small bank to eliminate the last traces of TVI." (Note: The catalogue of filters and other interference-control devices can be obtained from Electronic Specialists, Inc., 171 South Main St., Natick, Massachusetts 01760.)

Curt Powell, WB4WAA — "I had a similar RFI problem with my Montgomery Ward microwave oven Model KSA-8192A." Curt describes a frustrating experience he had with the company. He wrote the FCC about the problem, and received a reply by telephone. The FCC told him that the pulsed power supply energy was creating interference that was radiating down the power cable and being conducted back into the line by the con-

> *"The FCC may have received brickbats from hams, but here they are strongly on the side of the consumer!"*

nection. They informed Curt that the oven had been built by Sharp, and that Sharp was sending "a new transformer." Eventually the replacement part arrived, was placed in the oven by the Montgomery Ward Service Department, and the problem was solved. As Curt says, "The squeaking wheel gets the grease!" Obviously, the FCC had called both Sharp and Montgomery Ward over this problem. A follow-up letter was sent to Curt by the FCC.

the letter from the FCC

The letter Curt received was from Charles M. Cobbs, Acting Chief, Equipment Authorization Branch, Federal Communications Commission, Office of Science and Technology, Box 429, Columbia, Maryland 21045. The letter said, in part:

"We are very interested in the outcome of cases similar to yours. However, at the time of Type Approval of the model oven you have purchased, a line-conducted specification was not in effect for miscellaneous equipment, as microwave ovens are defined by our rules.

"In an effort to alleviate the problem, we are writing to the manufacturer of your oven and asking them to contact you directly and report to us their solution and specific action they used to bring the oven into a proper operating condition. Please inform us if nothing has been done within a reasonable time." (signed) Charles M. Cobbs

hurray for the FCC

The FCC may have received brickbats in the past from hams, but here's an instance in which they are strongly on the side of the consumer! And the Goldwater RFI bill should make the situation a lot better in the future than it has been in the past.

I suggest that if you contact the FCC regarding RFI from your microwave oven or other home equipment, you specify the name, model, and serial number of the equipment and the manufacturer and purchase date. It seems that when the FCC contacts some of these manufacturers, something happens.

the ground plane versus the dipole

In my October column I commented that tests I'd run comparing a ground plane antenna and a dipole seemed to show that the dipole *outperformed* the *vertical both for* transmission and reception. That conclusion brought some interesting comments from the readers:

Harry Hyder, W7IV — "In the late '60s I had both a ground plane vertical and an inverted-V. I was thus able to make direct comparisons. The vertical was a Hustler 4BTV, roof mounted, with four resonant radials, drooping slightly. The inverted-V was 35 feet at the apex and about 20 feet off the ground at the ends.

"At that time, I worked ZS1A frequently on 40-meter CW. I always heard him better on the inverted-V because of the lower noise level. On transmit it was a different story. He always heard me on the vertical (conditions permitting) but when I switched to the dipole my signal was always much weaker and I frequently dropped out completely . . .

"A friend of mine, W7IR, made the first 160-meter WAC from W7-land. He has a 90-foot vertical and a 160-meter inverted-V with the apex at 65 feet. The dipole is better for reception but is useless for DX transmission; the vertical outperforms the V by a large margin."

Warren Amfhar, WØWL — "From this writer's experience, when the ZLs are coming through on 80 meters during the early hours and a switchable vertical-to-horizontal antenna is used, the horizontal works better. Why? Because the signal is usually coming via multiple-hop, high-angle propagation."

Alan Bloom, N1AL — "Aren't you comparing apples and oranges? The radials on a ground plane 12 feet above ground do not act like an infinite ground. This antenna has a vertical radiation pattern similar to that of a dipole unless it has an "infinite" high-conduction ground at its base. Also, could the greater noise on the ground plane have something to do with its location nearby to house wiring? My own pet antenna is a high and long end-fed wire with lots of counterpoise wires. I used to have fun working Europe on 80 CW with my HW-16 with my "invisible" (No. 28 wire) 200-foot end-fed wire."

Roy Lewallen, W7EL — "In an experiment similar to yours, comparing a ground plane antenna with its bottom at 35 feet to an inverted-V dipole at about 40 feet, I found no direction or distance for which the ground plane did as well as the inverted-V on 20 meters.

"On 40 meters, a ground-mounted vertical with a good radial system approximately equals an inverted-V with its apex at 40 feet for signals beyond 500 to 1000 miles; closer stations are better on the inverted-V.

"When comparing vertical and horizontal antennas, signals must be monitored for some time to get accurate results. The antennas may in fact take turns being better than each other by 30 dB, because of polarization rotation of the received signals. On 40 meters, a typical period of rotation is from about one-half minute to a few minutes. The frequently heard question, 'Now I'm trying antenna number one. Now I'm using antenna number two. Which is better?' can

lead to results 30 dB in favor of one antenna at one moment, and equally in favor of the other a minute later!

"The effect of trees on vertically polarized signals can be significant. I measured a 10-dB loss through a small stand of fir, pine, and cedar trees which were roughly a quarter-wavelength high. A number of experiments left me convinced that the trees were indeed the culprit.

"In a few years the 24-MHz band will be dead . . . Let's use the band before this happens."

"The absorption of low-angle, vertically polarized signals is well known. One of the best illustrations appears in the Canadian Department of Communications CRC Report 1255 (unfortunately out of print). It clearly shows that absorption is a function of ground conductivity for some distance from the antenna, and it is not affected by the ground radial system as commonly thought (unless, perhaps, the radials extend for many wavelengths from the antenna). Since ground conductivity improves as frequency decreases, because of greater skin depth there is less absorption at lower frequencies. I suspect but haven't confirmed that a vertical is superior at and below 3.5 MHz. When compared with dipoles at heights at which the average Amateur is liable to put them, the vertical is almost certainly superior at the lower frequencies."

W.B. Pretchtl, W3KO — "I must take issue with the statement that it is difficult to devise a better antenna than a simple dipole. Putting aside the usual trials and tribulations accompanying a new antenna, I ended up with a rectangular loop, 40 feet on top and bottom and 27 feet on each side. The bottom is 7 feet above ground level and center-fed with a 4-to-1 balun and 50-ohm line. The VSWR on 40, 20, and 10 meters is better than 1.2:1. Its performance far exceeds expectations."

W3KO and N5CIE (XYL) point out that the loop functions well with low SWR on three bands. This is certainly a simple tribander antenna that doesn't require tuning and costs very little.

So there you are! It seems that the jury is still out as far as a meaningful decision between the horizontal dipole and ground plane vertical goes. I am interested in readers' comments on their comparison of the two antennas. Join in the fun!

18 and 24 MHz

Isn't the 10-MHz band wonderful? Plenty of stations to work, no contests to jar the nerves, and lots of good DX showing up every day. Now it's time to turn our attention to the 18-MHz and 24-MHz bands. I have monitored these bands and notice plenty of European Amateur activity, plus a few signals from South America and the Republic of South Africa. As of this writing (late November), no signals from "down under" or Asia have appeared.

Both bands seem nearly empty in comparison to other services. A few RTTY signals, one or two commercial CW stations — and that's all! The spectrum space seems wasted and I suggest it could be put to better use by permitting the U.S. hams temporary authority to use the bands before they become useless. Remember, the sunspot cycle is declining and in a few years the 24-MHz band will be dead for long-distance communications. It would be nice to use the band before that happens.

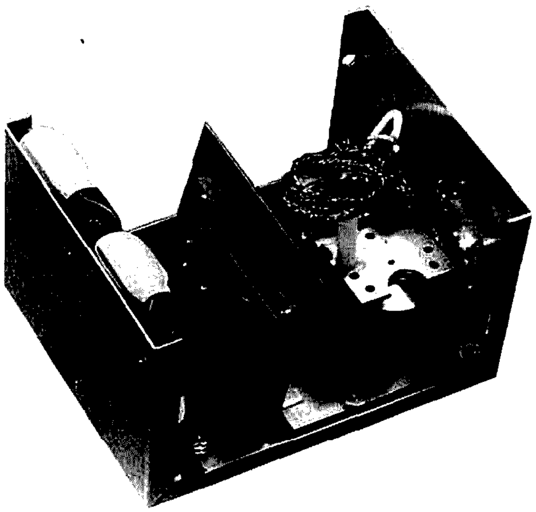
ham radio

Multiple-line transformers improve accuracy making possible realistic phased-array impedance measurements

a precision noise bridge

An rf impedance bridge is an important tool for anyone with more than a casual interest in measuring complex impedances. One typical application for such a bridge is in the design of a phased-array antenna.

Solid-state technology has made it possible to move the rf impedance bridge out of the laboratory and into the average ham shack, but, unfortunately, laboratory accuracy cannot usually be achieved with most of the available units. Although readings within 10 percent are acceptable for most Amateur applications, much better accuracy and discrimination are necessary when taking measurements of antenna-array mutual impedances. We must in fact be able to accurately determine resistive components of impedances which may change as little as only 3 or 4 ohms when a nearby antenna element is mutually coupled. Obviously, most noise bridges can't do the job.



Internal view of noise bridge showing coupler and PC board placement.

This article explains how it is possible to achieve the needed accuracy in the range of the hf Amateur bands. A multiple-line, distributed-impedance transformer is the critical bridge component. It is not very susceptible to fabrication variations, and it maintains a single calibration over a wide frequency range. With care in adjustment and good construction practice, it is possible to achieve an accuracy of 3 percent over most of the range — with even better results at 14 MHz and lower.

background

In an excellent article on noise bridges,[1] W6BXI and W6NKU contributed two major innovations for improving noise bridge accuracy: compensating for bridge circuit strays by adjusting inductance in one of the bridge secondary arms; and equalizing primary-to-secondary interactive effects by the addition of a dummy primary wire.

The first idea is one of those insights that seem so obvious after someone else has pointed it out: One wonders why the C_p calibration "rotation" problem at high frequency ever seemed so difficult. The second suggestion is logical, but I could confirm it only empirically. After winding a dozen different kinds of

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three-wire and four-wire toroid transformers and comparing results, I leaned toward the four-wire version. I became convinced that control of interactive effects between the primary and secondary was a key factor in achieving the accuracy I wanted.

I used the improvements and suggestions in the referenced article in three of my own bridges, with good results similar to those achieved by the authors: a spread of 2 to 3 ohms (resistance and/or reactance) for a range of accurately known impedances. But further improvement in accuracy was still needed.

W6BXI, in private correspondence, told me that he'd had no particular difficulties with the bridge transformer, either in reproducing it or with circuit adjustment. I wasn't as lucky. But he did discuss a method of matching and balancing the bridge transformer to the noise source. I tried an emitter follower that provided a low-impedance source for the noise amplifier. I also tried a balun toroid transformer, to take advantage of balanced drive to the double-ended bridge circuit from the single-ended noise source. Neither approach worked.

I have probably wound a hundred or more different configurations of the toroid transformer used in this bridge, in an effort to get a better understanding of this key component in the bridge. One possible solution lay in trying to get maximum inductive coupling with minimum capacitance between primary and secondary. So I wound a more classic type of transformer, with the primary on one side of the toroid and a single-wire, center-tapped secondary on the other. Capacitive coupling between windings was very low — 3 pF — but the transformer behaved poorly in the bridge. Calibration barely held over a single Amateur band, and there was no correspondence to the real values of the reference arm of the bridge. Next I wound a second transformer, also with the windings on opposite sides of the toroid, except that now the secondary was a bifilar winding. The bridge worked, and calibration held from 1.8 to 30 MHz. But the noise signal was too low at the high-frequency end, indicating insufficient primary-to-secondary coupling, resulting in poorly defined nulls. Different core materials did not help. This experiment showed me that the required bridge coupler was no ordinary transformer, but something akin to a distributed impedance transmission line.

Most of the trifilar and quadrafilar toroid transformers I tried had been wound with the three or four wires kept carefully parallel. I wound some with the wires twisted together at two or three turns per inch. A stranded bundle of approximately 9 inches (23 cm) was wound on the core. I decided to try connecting up a strand in the usual way, but as a large single-turn loop, without winding it on a toroid core. This worked excellently, without any of the puzzling

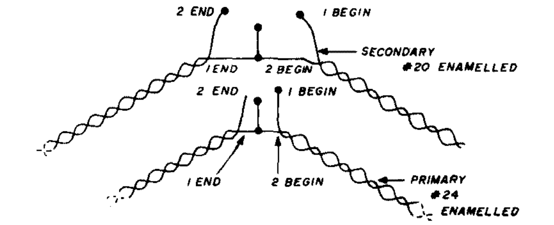

fig. 1. Bridge coupler twisted wire pairs interconnection diagram.

anomalies of the toroid versions. Noise signal was as high as with a toroid, and calibration held constant from 1.8 to 30 MHz. Also, the reference potentiometer settings closely coincided with the value being measured. With this approach there was no question about the four-wire line being superior for holding calibration over a wide range, and my results were reproducible: Reproductions of these lines always produced the same results. I concluded that the toroid core had been contributing most of the anomalies — and that it was not at all essential to this bridge component.

Having eliminated the toroid, I next worked on optimizing line length, wire size(s), and insulation thickness, and on a method of combining the individual wires into a line. The matter of whether adjacent or alternate wires of a single twisted line should be used in connecting up the primary and secondary circuits was also investigated. I tried dozens of variations: No. 24 (0.5 mm) hookup wire with PVC insulation, enameled wire sizes from No. 18 (1.0 mm) to No. 32 (0.2 mm), including mixed sizes, lengths from 4 inches (10 cm) to 16 inches (40 cm), single twisted and compound twisted, and so on. This is what I learned:

1. Close coupling between primary and secondary is absolutely required. This means using enameled wire, twisted as tight as possible.

2. Line lengths between 10 and 12 inches (20 and 30 cm) are best: If too short, noise signal is too low; if too long, a large amount of compensation is required on one secondary half for R_p adjustment, which is accompanied by a pronounced C_p calibration shift between 21 and 30 MHz.

3. Compound twist gives the most consistent results (primary and secondary wires are first twisted independently, then these are twisted into a single strand). All four wires can be twisted simultaneously, but they would then have to be interconnected with attention to whether alternate or adjacent wires are being used as the respective primary and secondary. See **fig. 1** for winding interconnection detail.

4. Enameled wire size No. 20 (0.8 mm) seems best. Smaller wire sizes are more difficult to connect,

adjust, and compensate because of their fragility. Heavier wire, though effective, resulted in a line that became too stiff to work with. I found a good compromise to be No. 20 (0.8 mm) for the secondary and No. 24 (0.5 mm) for the primary. These lines can be looped into multi-turn coils, to conserve space, with no effect on calibration. In the single-turn calibration no ill effects occur unless the loop is tightly collapsed on itself, and then the most noticeable effect is a resistance calibration shift at low frequencies.

These coupling devices act like distributed-impedance transmission lines. The circuit impedance appears to be low, as it is virtually immune to stray capacitances. As with transmission-line baluns, however, the device displays sensitivity to lead dress of the ends of the lines. Unfortunately, it is not physically possible to keep these ends apart. One end of each pair has to be connected to each other, and the ends connected to the bridge arms must be as short as possible. If, indeed, this is a multiple-wire transmission line, theoretical calculations would point toward a low characteristic impedance. The effect of the re-entrant connection is difficult to assess. I would welcome letters from readers who are able to show the mathematics. The most puzzling aspect, to me, is that a line so short (in terms of wavelength) should be effective at very low frequencies. I have used a bridge *calibrated* for frequencies between 3.5 and 30 MHz to as low as 150 kHz. The calibration was unchanged and the noise signal level was as high as at 3.5 MHz.

scale extensions/expansion

The article by W6BXI and W6NKU and subsequently published letter[2] give a technique for measuring impedances beyond the basic range of the bridge by using series or parallel resistances inserted at the **Unknown** terminal. Though this approach is quite effective, the user is warned that, as with any scale multiplier, inaccuracies in the basic range calibrations are also multiplied. Improvement of the base range accuracy should make these range extenders more trustworthy — and using the minimum multiplier possible. For antenna measurements the series extender finds the greatest use; I have several, ranging from 10 to 100 ohms.

For some applications, like mine, there is a need to expand the resistance scale to be able to discriminate between small changes. Since the R_p range of interest was between 27 and 77 ohms, a 50-ohm potentiometer in series with a 27-ohm fixed resistor was used in the reference side of one of my bridge models. This affords nearly 270 degrees of scale rotation for a translated 50-ohm range. To minimize strays, the potentiometer connections should be as shown in **fig. 2**.

transformation equations

A series combination of resistance and reactance can always be found that exhibits the same equivalent impedance as any given parallel combination of resistance and reactance.

The transformations give the relationship between the elements of the series and parallel networks (shown below) when the driving-point impedances are equal. The equations for the respective impedances are:

$$Z_p = \frac{R_p(jX_p)}{R_p + jX_p}$$

$$Z_s = R_s + jX_s$$

Equating the real and imaginary parts of both expressions for network equivalency yields:

$$R_s = \frac{R_p X_p^2}{R_p^2 + X_p^2}$$

$$X_s = \frac{R_p^2 X_p}{R_p^2 + X_p^2}$$

Transforming from series to parallel:

$$R_p = R_s + X_s^2/R_s$$
$$X_p = X_s + R_s^2/X_s$$

other modifications

The other major changes from prior circuits were to go to a fixed capacitor in the reference arm of the bridge and to place the variable capacitor, C_p, across the **Unknown** terminal. Operation of the bridge remains the same, except that the meanings of the direction of rotation from "0" on the calibration scale become reversed. That is, an indicated null which *increased* the capacitance above its value at "0" means the measured impedance has an *inductive* reactance component. Conversely, a decrease in capacitance from "0" indicates a capacitive reactance. This change ensures that the reactive component being measured is in direct correspondence with that of the unknown impedance. The difference is a subtle one but, in effect, at nulled conditions the reference arm is always "seeing" the identical situation, the fixed capacitance across the potentiometer.

adjustment

Two adjustment procedures are necessary, both requiring a bit of care and patience. One is done at low frequency to enable the unknown load resistance to track with the reference potentiometer's nulled resistance. The second adjustment is done at the high frequency end of the range to enable the load resistance to continue to track with the potentiometer and to reduce the C_p calibration shift. The adjustments are performed in the order given and must be iterated since they are interdependent. They are best done with a good commercial load termination (of minimum reactance), though 50 to 100 ohm quarter-

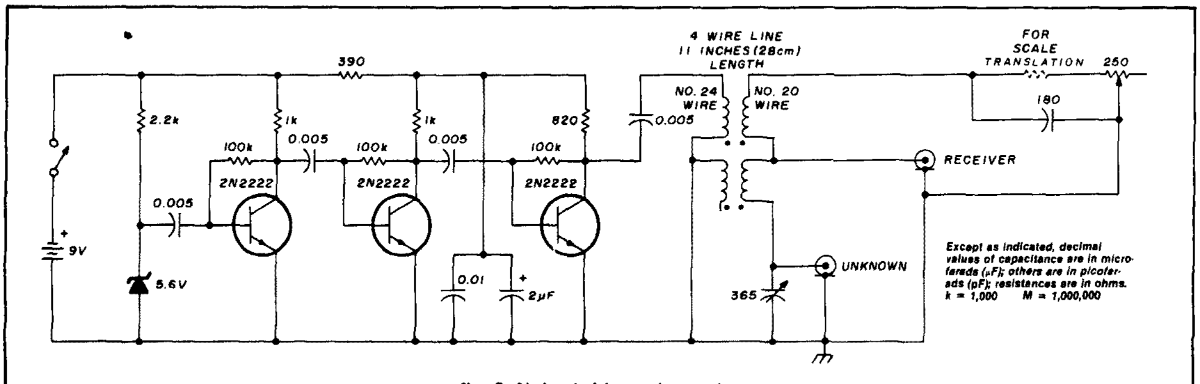

fig. 2. Noise bridge schematic.

watt carbon resistors connected by short leads are a second best alternative. These adjustments ensure that the calibration remains constant over the bandwidth and provides an accurate R_p readout. I have found multi-tester DWM's to be remarkably accurate ohmmeters and useful in measuring the calibration load resistances and the nulled potentiometer resistance.

The low-frequency adjustment is better described than explained. The adjustment cancels inductive and capacitive strays, and after building six bridge units, only two with identical layouts, I am confident of the procedure:

low-frequency adjustment

1. Set the null-detecting receiver to a low frequency (I used 3.5 MHz) and null the bridge with a known resistive load of 50 to 100 ohms. The R_p and C_p controls are interdependent; repeat for best possible null.

2. Turn off the noise source and remove the load and null detector from their terminals. Measure the potentiometer resistance without disturbing its setting (measure the resistance seen at either the **Unknown** or **Detector** terminals). Measure the load resistance.

3. If there is agreement to within 0.1 ohm go to the high-frequency adjustment procedure. More likely, the agreement will be close but not good enough. With the bridge returned to operation as in step 1, gently bend the bridge line to a new position in any direction by a few degrees. The potentiometer null will change; renull and note whether the potentiometer resistance more closely agrees with the load. If it does not, bend the line in another direction, possibly opposite to the direction of the first trial. If after a few trials it is found that no direction of movement will cause good enough correspondence of readings, go on to the high-frequency compensation procedure. The two procedures are interdependent and several iterations will be necessary.

high-frequency adjustment

This adjustment equalizes any residual imbalance of inductance that may exist in the bridge. The adjustment is made at the high-frequency end of the range because this is where the influence of an imbalance is most noticeable. It consists of adding inductance (offset) to the secondary arm found to require compensation. Which arm this is depends upon the direction the R_p null must be shifted. I have found it always to be required in the reference arm connection if the noise signal is inserted to the primary connection at the same end of the line.

1. Using the same pure resistance load as in the low-frequency adjustment, null the bridge carefully at 30 MHz. Turn off the bridge, remove the load and detector, and compare the potentiometer resistance with that of the load. If this is the same you had at 3.5 MHz you're done. More likely, compensation will be necessary.

2. Add an inch (2.5 cm) of the same size wire used for the secondary to the secondary connection at the reference bridge arm. Returning the bridge to operation, if the R_p null resistance is higher than the resistance being measured and now goes higher still, remove the wire and add it in on the other secondary arm of the coupler. If it now goes lower, reduce the amount of added wire until the R_p null resistance matches the termination.

3. In making this adjustment be careful not to disturb the position of the line found for the low-frequency adjustment. In any case, recheck the low-frequency adjustment after completing this procedure. Several repetitions of the two adjustment procedures may be necessary. If more than an inch of secondary-size is required, you may encounter an unacceptable C_p calibration shift at 30 MHz, although R_p tracks over the whole range.

These two adjustment procedures force tracking of R_p at widely separate frequencies. But this does not guarantee flat tracking at frequencies in between, so after completing these adjustments check the calibration in the middle of the range, such as 14 MHz. R_p calibration should remain well aligned and C_p shift, compared with the 3.5-MHz position (the center, or "0" of this scale), minimal (less than 1 pF). If all is well, now check with other values of resistance. If you cannot obtain commercial terminations, quarter-watt carbon resistors of the deposited type are quite good, but connect them to the unknown terminal with as short a lead length as possible. In spite of all your best efforts, you will see more C_p inductive shift at 30 MHz with these. If you used a 250-ohm potentiometer, expect to see more absolute R_p shift at 30 MHz than found at 3.5 MHz for resistance at the upper end of this range. My bridge reads 1 ohm low with a 237-ohm load at 30 MHz. At the low end of the range, using a 6-ohm resistor, the R_p reading is within 0.1 to 0.2 ohm over the whole frequency range, with one bridge reading on the high side at 30 MHz and the other on the low side. A DVM checked against a laboratory standard was used for a resistance calibration standard.

"An rf impedance bridge is an important tool... in the design of a phased-array antenna."

If the potentiometer is not tracking well with frequency, the circuit has layout problems. Here are three general areas to review:

1. Ground loops in the circuitry (multiple connections to ground for the same circuit path).

2. Too much inductance has been allowed to creep into the connections in known or unknown arms of the bridge. One clue is a continuous shift of C_p center as you step up in frequency bands from 3.5 to 30 MHz.

3. The noise source ground connection to the bridge has too much impedance. This is indicated by a pronounced shift in C_p center position at 30 MHz, with little or none at 21 MHz. The effect is very different from the prior case, in which there is gradual shift. Note that this same effect also arises from a requirement for a large amount of compensation at high frequency. *Still*, this is related; the amount of compensation is an indicator of stray reactance in the circuit.

circuit hints

Avoid creating ground loops. Have all ground returns going to a common point; this point is best located at some point midway between the circuit connections for the known and unknown bridge arms. Calibration accuracy to 30 MHz requires extreme care here. For example, in the last two units I built the detector coax terminal is isolated from the chassis. Coax (RG-174) is used to continue this line directly to the bridge connections, including ground. The bridge circuit and components are mounted on a printed-circuit board using copper foil on the circuit side. The chassis screw mountings for this board are insulated from the foil. The only connection with the chassis and the copper foil occurs via the SO-239 **Unknown** coax terminal.

Any stray reactance in this bridge creates calibration shift problems, particularly at 30 MHz, and if severe, even at 14 MHz. I found that liberal use of flat, braided shield (such as is used on RG-58/59 coax) for all bridge connections is indicated. In this regard, the miniature 365-pF variable capacitor recommended in the referenced article has too much internal inductance. The small Japanese air variable capacitors built on heavy aluminum frames have been found to be excellent and quite linear.

As a general rule disproportionate shifts in calibration with increasing frequency are indicative of stray capacitance problems, while linear shifts are due to unwanted inductance.

As for the potentiometers, Mil. Spec. linear-resistance carbon pots are recommended. I found Clarostat Type 53C1, S-taper to be best. These are available in resistances as low as 50 ohms and are very linear.

As for the noise source, obtain several samples of the zener diode for the *noise amplifier. Select the one* with the highest noise output over the frequency range, emphasizing 30 MHz. Some diodes are so quiet there will be difficulty obtaining enough signal. Isolate the noise amplifier and its batteries from all grounds except the connection to the common ground at the bridge circuit. In one of my units, flat braid for this ground return solved a stubborn residual C_p shift problem at 30 MHz.

Unless you have a compelling need to carry a bridge in your hip pocket, don't try to crowd this device into a snuff box. My units are housed in a standard aluminum two-piece box 5 × 4 × 3 inches (13 × 10 × 8 cm). This is small enough, gives you space to work in, and keeps down strays. With the extra

space available another 9-volt transistor battery can be paralleled for extended operation.

C_p dial calibration

With R_p calibration done, you are ready for calibration of the C_p dial. Assuming you have found minimal or no C_p shift of the "0" position with a 50-ohm termination at 30 MHz, I recommend that 3.5 MHz be used for this calibration, since it minimizes the effect of lead inductance errors.

In the referenced article[1] the authors suggest the following procedures: Calibrate the variable capacitor against a standard before connecting it to the bridge circuit. Or calibrate the capacitor, with bridge operating, using known values of capacitors in parallel with a resistance at the unknown terminal (50 ohms suggested). After completing the capacitive portion of the dial, temporarily disconnect the 180-pF bridge centering capacitor and work backward toward "0" for the inductive portion of the dial.

The first method is the preferred one. But since I had no capacitor standard available, I used a variation of the second method: After calibrating the capacitive portion of the dial, connect a 50-ohm quarter-wavelength coax cable cut for 3.8 MHz to the **Unknown** terminal. At the far end connect a 39 or 68 ohm resistance with minimum lead length. Find the frequency which nulls this resistance while the C_p dial is set at "0." Note the frequency and remove the resistor. This is the exact frequency at which your cable is a quarter-wavelength. Now use the same known capacitors in parallel with 50 ohms at the end of this cable to calibrate the inductive side of the dial. The quarter-wavelength coax causes these capacitors to look "negative," that is, inductive by a nearly equal amount. Since these are *parallel* circuits *in* series with the coax, a small correction is necessary which increases with increasing capacitance. Assuming a 50-ohm coax with a 50-ohm termination at 3.8 MHz, use **table 1** correction points (interpolating for intermediate calibration points).

table 1. Parallel-series-parallel correction for use in calibration of inductive side of C_p dial.

capacitance (pF) at end of coax	capacitance (pF) at terminal
50	49.82
75	69.51
100	98.60
120	117.59
130	126.94
140	136.20
150	145.34
160	154.37
170	163.28
180	172.06

upper frequency limitations

Frequencies higher than about 20 MHz begin to present problems for any circuit measurement device. This is why it is not easy to achieve a constant calibration with good accuracy as we go up in frequency. Here we enter a realm in which a resistor, a capacitor, or a coil can no longer be thought of as discrete elements; indeed, each can be a combination of all three. Since this also applies to the bridge circuit, calibration correction and circuit adjustment schemes become necessary. Reactance is involved, and so these compensations are frequency dependent. This means calibration is meaningful over relatively small frequency ranges. I have checked my noise bridge as high as 100 MHz and found that, while deep nulls can still be detected, the calibration shifts considerably.

low resistance limitations

Measurement of very low R_p circuits (below 5 ohms) poses two problems for this bridge: The low R_p, coming quite close to a short circuit, reduces the C_p null sensitivity, making it more difficult to determine balance. At the same time, any accompanying reactance, since we are measuring the *parallel* equivalent of the circuit, results in large excursions of the C_p dial for relatively small reactances.

As R_p approaches zero, bridge circuit strays become more significant. Do not depend upon any measurements made with this type of bridge with the R_p dial setting near zero, since many potentiometers do not actually reach zero resistance at their mechanical stop. I have heard of attempting to determine a quarter-wavelength of coax this way. This bridge can do that and more, but not in this way. If measurement of very low impedances is an objective, use a series extender, or consider a series-type bridge.

detector considerations

One of the reasons the noise bridge is a relatively simple circuit is that no null detector is included. For this an ordinary receiver is used. This can be the station receiver, and, in these days of accurate receiver frequency readout, it can be a frequency standard as well. For purposes of bridge calibration an Amateur-band-only receiver is adequate, but for most measurement needs a general-coverage receiver is better. The presence of an S-meter is helpful but not necessary.

When making measurements on antennas, a battery-operated receiver is convenient. I use an inexpensive all-band portable and a transistor crystal oscillator for marker frequencies. Since the noise signal is considerable, (S9 +20dB, off null), receiver sensitivity is no great consideration; too much can lead to difficulty in finding a null because of receiver

AVC. The sharpness and depth of nulls provided by this bridge requires getting used to!

I have used as much as 30 feet of coax between the bridge detector terminal and the receiver with no effect upon calibration.

An fm receiver will not work well with this device, as the noise is almost entirely a-m. If the receiver has AFC it won't work at all, as AFC always shifts the receiver local oscillator off the null frequency.

applications

Before discussing applications, I want to mention the difference between parallel and series circuits. Each type can be transformed into the other. This bridge measures parallel-circuit equivalent values for *both real parallel and series* circuits. If a real parallel circuit consisting of a 50-ohm resistor and a 180-pF capacitance is measured, then that is what is read out, regardless of the frequency. But a series circuit of 47.8 ohms and 4078 pF at 3.8 MHz also will be read by the bridge as 50 ohms R_p and 180 pF C_p. This point must always be considered when using this bridge.

An excellent tutorial on bridge applications and calculations using complex algebra exists.[3]

coax cable measurements

If you used my method to calibrate the C_p dial, you have learned how to use this bridge to find very accurately the frequency for a quarter-wavelength of coax. If you have R_p, then taking the square root of the product of R_p and the resistive termination yields the characteristic impedance of this line. The ratio of the physical length of the line to the free-space quarter wavelength for the frequency measured is the velocity factor. If you were working with foam cable, you might be in for some surprises at this point. Neither of your calculations may agree very closely with the nominal values usually quoted. Characteristic impedance can vary as much as 10 percent, and velocity factor is seldom as high as the 0.82 usually given; 0.70 is more likely.

When using this method, choose a termination resistance different from the characteristic impedance, but one which will yield a transformed resistance within the range of the bridge. As a check on your work, try a resistance termination on the line equal to your calculation for characteristic impedance. You should find that the bridge, after being nulled for this resistance, maintains that null over a wide range of frequencies.

I have found that quarter-wavelength cables determined *via grid-dip methods are* 1 to 3 percent too long, when compared with my results. The grid-dip method introduces error because of the shorting link and the pickup coil.

antenna measurements

An obvious application for this bridge is the measurement of self and mutual impedances of antenna elements and matching adjustments. Measurement of mutual impedances must be done indirectly and involves fairly complex calculations. (This will be discussed at greater length in forthcoming articles.)

Self impedance measurement is straight-forward: Simply connect the antenna leads to the bridge and adjust for null. Remember to keep lead length the same as in the actual installation, and remember too that the values being read are the parallel-circuit equivalents.

A very nice application for an accurate bridge is measuring the impedance of antennas right in the shack, having previously obtained an accurate measurement of the feedline length and its characteristic impedance. Using a Smith chart or programmable calculator to rotate the measurement back to the antenna saves a lot of legwork.

One point particularly applicable to antenna measurements is, make doubly sure of all connections and joints. This bridge operates with noise power measured in microwatts. Poor connections, which do not show up in normal operations, even when driven QRP, will become evident. A few watts may temporarily "weld" poor connections; the bridge hasn't enough power to do that.

impedance transformers and networks

Another useful application for this bridge is measurement of the input and output impedances of transformers and networks. Remember that the readings are in parallel circuit form, and that the terminations may be in either form. It's sometimes easier to arrive at one than the other with available components.

references

1. Robert Hubbs, W6BXI, and Frank Doting, W6NKU, "Improvements to the RX Noise Bridge," *ham radio*, February, 1977, page 10.
2. Robert Hubbs, W6BXI, and Frank Doting, W6NKU, "Antenna Noise Bridge" (letter), *ham radio*, September, 1977, page 100.
3. Leonard Anderson, "Antenna Bridge Calculations," *ham radio*, May, 1978, page 34.

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ham radio

the ham notebook

DC to 400-Hz ac converter

There are many inexpensive transceivers on the market that require 400 Hz. After I bought an ARC 38A, I was confronted with the problem of how to obtain 400 Hz for the servos.

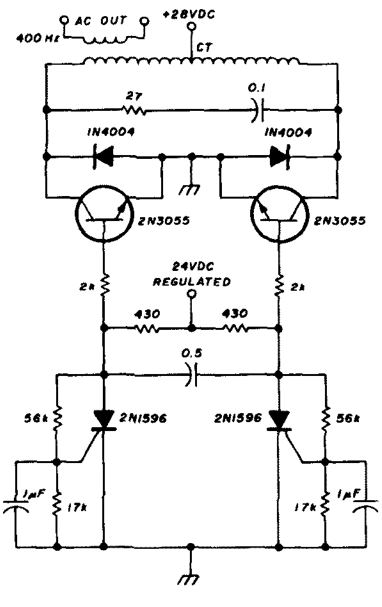


fig. 1. DC to 400-Hz ac converter.

Self-oscillating transformer circuits exist, however they need special transformer ratios to work and if loaded they like to change frequency or stop oscillating. A 555-oscillator chip with a driver and a push-pull output is another choice, but I wanted something simpler.

The circuit shown works well and is not too elaborate (**fig. 1**). It depends on inherent unbalance to start. Any small SCR will do.

In the oscillator[1] the voltage divider starts charging the capacitor until one SCR breaks down, initiating oscillation. The driver transistors (2N3055 or equivalent) drive the transformer with a square wave. Little power is dissipated in the transistors since they are mostly on or cut off. Diodes from collector to ground prevent a negative kickback voltage from appearing. 27 ohms and 0.1 μF are used to dampen spikes.

reference

1. W.B. McCartney and E.O. Uhrig, "Astable High Power Multivibrator," *IEEE 10;12*, pages 30-31.

Kurt A. Bittmann, WB2YVY

screen protection for the 5CX1500A

If anyone is considering building a linear amplifier around the 5CX1500A tube, he should not neglect to include screen protection circuitry. Because of the preponderance of grounded-grid triode amplifiers in the last few years, this important point is now only superficially mentioned in the Amateur literature, and I have not seen it incorporated in published circuits. It is all too easy, however, to exceed the screen rating if there is instability during testing, or by resonating the tank with light loading.

During the construction of a 4CX-1000A amplifier I came across an article on tetrode screen current in an old *QST* (David D. Meacham, W6EMD, "Understanding Tetrode Screen Current," July, 1961). It convinced me of the importance of the screen overcurrent relay, and I designed an appropriate circuit based on the author's ideas (see fig. 2.)

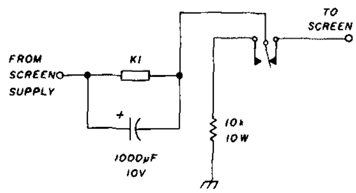


fig. 2. Screen current relay.

I selected a relay sensitive enough to pull in if the current exceeded 60 mA, and yet with resistance low enough not to impair screen regulation. In my case, this was a Japanese-made subminiature relay with a 6 volt, 80-ohm coil, but ITT also makes a line of similar relays. R1 must draw enough current to hold the relay on with the supply voltage in use. A relay requires far less current to hold it on than to pull it in.

During testing and tune-up, this relay has paid for itself many times over. Thanks to its quick action, my final has avoided the fate of irreversibly turning into a low-mu triode.

Vic Mozarowski, VE3AIA

solid-state CW T-R system

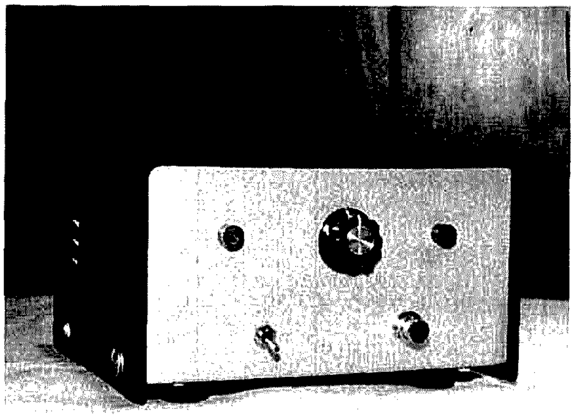
A changeover system suited for older equipment

For the Novice operator and the CW fan, there are many older transmitters and receivers that will still perform with the best of the modern transceivers. And, for many of us, the price is right. They take up a bit more room, and we may have to work on them to hold down the drift and keep the performance up. But that is part of the challenge we accept when we decide to use such classics as the Heath DXs, the Globes, the Hammerlund HQs, or the National NCs.

Perhaps one of the greatest inconveniences inherent to most of these otherwise fine designs, however, is the lack of an automatic or semi-automatic transmit-receive system. In fact, most of these rigs lack an antenna change-over relay: switching is done manually. It's necessary to turn the transmitter from standby to operate and mute the receiver as well. In some cases, you must also turn on the VFO. All this switching can get in the way of smooth operating.

CWX: tube style

Hams solved the basic problem of system switching long ago with a system called CWX, CW-operated relay. A CWX automatically switches all rig functions from receive to transmit with the first tap

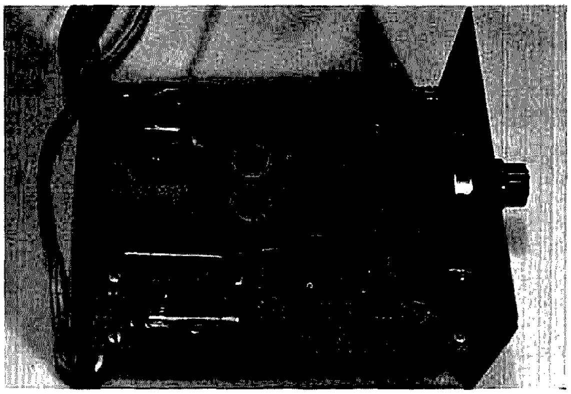


The front panel of SSCWX is simplicity itself. The top row consists of the ac power LED indicator, the timing potentiometer, and the transmitter-on LED indicator. In the bottom are the ac toggle switch and the key jack.

By L.B. Cebik, W4RNL, 5105 Holston Hills Road, Knoxville, Tennessee 37914

of the key. In most units, a delay circuit holds the rig in transmit mode between dots, letters, and words, but after a second or so the relay returns everything to receive.

Most of the good CWX circuits used tubes and plate relays. Fig. 1 shows a typical circuit, used at W4RNL for many years in conjunction with a DX-60A transmitter and a 2-B receiver. It is a modified version of a circuit that originally appeared in *QST* in February of 1966. The modification consists of eliminating one relay and tube section (which followed the keying) and replacing it with the pair of diodes near the key jack. With blocked-grid keying, this system is effective.



The interior layout of SSCWX. In the left rear of the perf board are the power supply components and the function-changing relay. In the front right are (from bottom to top) the opto-isolator, the voltage comparator, and the switching transistors (mounted in an 8-pin DIP socket). Layout is non-critical if all leads are well bypassed as they enter the cabinet.

This old circuit will still work well with almost any miniature triode, and it has some advantages that are worth preserving. Note that the key line to the transmitter goes through spare contacts on the antenna relay (a Dow Key rf relay, in this case) so that the transmitter cannot generate rf until the antenna changes over. Spare contacts on the plate relay control the grid-block voltage to the VFO, holding the unit on during the entire transmit cycle. This feature eliminates chirp when the VFO frequency is multiplied to 20, 15, and 10 meters (a problem common to many inexpensive VFOs). The potentiometer permits adjustment of the delay time to suit varying code speeds. And finally, the unit is compact and self-powered, making it easy to fit either on the table or in a spare corner of transmitter or receiver.

The unit does use tubes, however, which have grown expensive in this silicon age. And plate relays are growing harder to find, although 12-volt relays are still plentiful. As a result, when one of the students in my license-upgrade class asked about an antenna change-over device for his DX-60, I wondered if I could preserve all the good qualities of the old CW in a solid-state device using easily obtainable components. SSCWX is the result.

CWX: solid state

The solid-state version of CWX is designed for blocked-grid keying and a variety of auxiliary func-

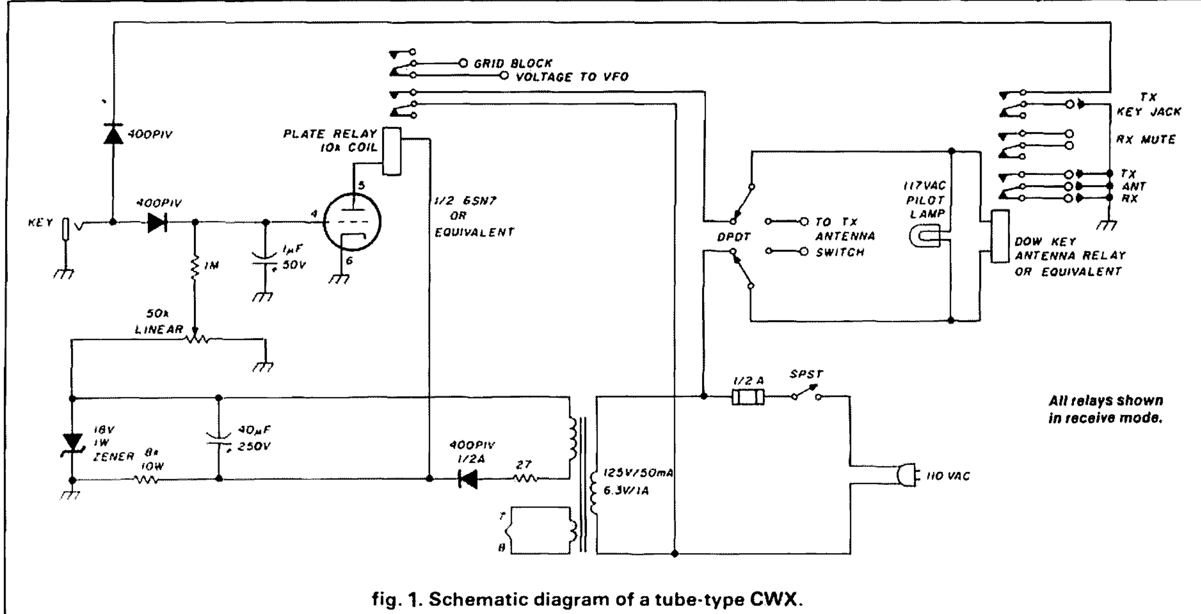


fig. 1. Schematic diagram of a tube-type CWX.

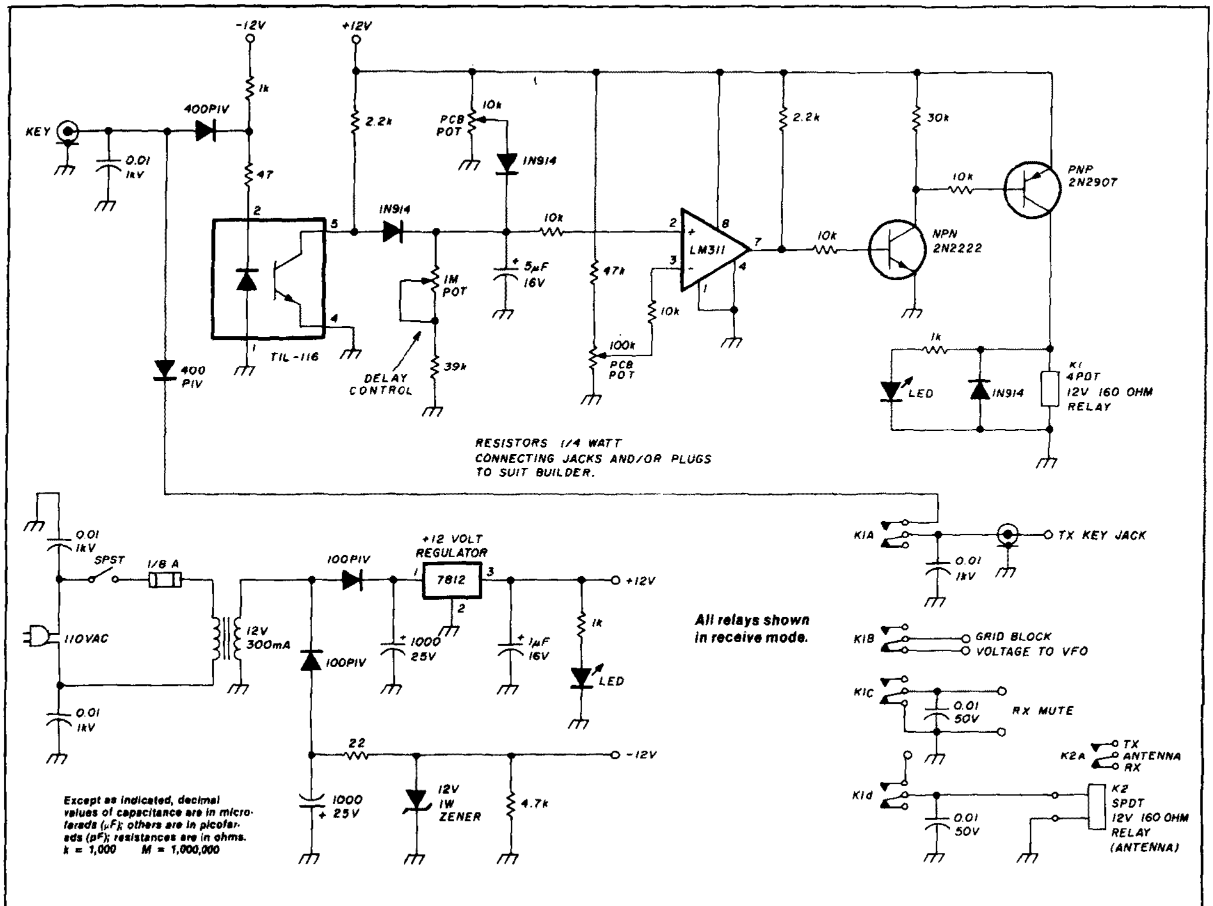


fig. 2. Schematic diagram of SSCWX.

tions, such as receiver muting, VFO control, and rf generation control. Given the variety of receivers available, each with a different muting system, control by relay offers more versatility than does solid-state switching. (Notes on customizing the design and expanding its use to other relay-switching applications appear at the end of this article.)

In older transmitters keying is usually either cathode (positive voltage, heavy current) or blocked grid (negative voltage, low current). In general, you should stay away from cathode keying, because the current will eventually wear key contacts or cause damage to the switching circuits of many keyers. Cathode-keyed transmitters can be modified to a blocked-grid system, but notes on altering SSCWX for positive keying also appear at the end of the article.

circuit operation

The SSCWX circuit performs in the following sequence (see **fig. 2**). The TIL-116 is an opto-isolator used to separate the key line from the switching circuits. The diode is lit with negative voltage, corresponding to the grid block voltage on the key line. Key down turns off the diode and the transistor stops conducting, sending the collector voltage to the plus terminal of the LM311 voltage comparator and to the 5-µF timing capacitor. The 1N914 diode in this line keeps the capacitor from discharging through the opto-isolator when the key goes up, and the TIL-116 once more conducts.

Between the opto-isolator and the voltage comparator is the timing circuit. The electrolytic capacitor charges quickly but discharges slowly through the 1-megohm timing potentiometer and the 39k fixed resistor. The fixed resistor controls the shortest time delay. The combination of the 47k resistor and 100k pot in the minus comparator input line controls the longest delay time, because the LM311 will change state whenever the voltages at its inputs pass each other. About 7 volts to pin 3 of the LM311 yields al-

most 3 seconds of delay – more than enough for most purposes.

The turn-on time is a function of the timing capacitor and the 2.2k collector resistor of the TIL-116. A single short dit will not yield the full delay time, without the supplementary charging circuit composed of the 10k pot and the second diode. By holding the timing capacitor at a level about 1 volt less than the comparator negative input line, the delay time is consistent for single key taps and actual CW transmissions.

The voltage comparator provides a full output swing whenever the capacitor voltage moves above or below the negative input level, going high when the plus line exceeds the minus and low otherwise. This swing triggers the two-transistor switch to the relay, keying it with definite action. The definiteness of the wave shape controlling the relay is the primary reason for using the comparator to isolate the timing circuit from the relay. Timing circuits tied directly to the 2N2222 base drive both transistors through the linear range, and the relay closes and opens as the voltage and current pass the threshold level for its coil. Neither of these phenomena is desirable. Since transistors draw base current (while the older tube circuit did not draw grid current), the devices are easily overloaded during long delay periods. In addition, many inexpensive low-voltage relays do not open cleanly as the voltage drops slowly, creating some sparking and consequent contact damage. Clean switching to the relay and to the controlling transistors is necessary, and the LM311 (or any similar voltage comparator having a fairly high input impedance and able to work from a single 12-volt line) performs admirably.

The 12-volt relay controls four circuits: receiver mute, VFO, antenna relay, and (possibly) the key line. Since Dow Key relays have grown very expensive, most hams have turned to ordinary heavy-duty contact relays for antenna switching. If the internal contact lines are fairly short, and if the relay is well shielded in a metal box with short leads to the three jacks (antenna, transmitter, and receiver), ordinary units perform well and do not increase SWR. If you have such a relay, however, do not use the extra contacts for key lines, muting or other such purposes, as they will be in a heavy rf field. Use a different relay. The circuit compromise in the SSCWX design does not guarantee that no rf will emerge from the transmitter until the antenna relay has changed over, but it does guarantee that malfunctions in the CWX unit itself will not permit you to pump rf into the receiver. **Fig.** 3 shows the relay circuit for use with Dow Key and similar rf-shielded relays.

Note that the VFO line shows the grid block voltage being controlled by the CWX relay. This represents a line from the VFO to the CWX to ensure that the grid blocking voltage is off for the entire transmit cycle. Modification for other schemes that help control VFO stability is simple. Some hams prefer to add voltage-controlled diodes to VFOs to shift them off frequency when the key is up, in which case the relay can handle the control voltage instead of the grid block voltage. Whatever the system, I do recommend that you use something to hold a tube-type VFO on for at least the entire transmit period, since the transition from cut-off to full plate current adversely affects short-term VFO stability, and signal quality as well.

Power for the SSCWX comes from a very small supply. The 7812 positive voltage regulator needs no heatsink in this application, but adding one cannot hurt, especially if the unit provides power to other accessories. If so, increase the transformer rating as well, since the entire unit, with two 75-mA relays, requires over two-thirds of the rated 300 mA from the secondary. The requirements for negative voltage are small, and the Zener may not even be required, although it does make the voltage predictable.

construction and adjustment

The entire SSCWX unit fits on a small piece of perf board, cut to fit a standard 5-1/4 × 3 × 5-7/8 inch (13.3 × 7.6 × 14.9 cm) box, with room to spare. The relay and the power transformer are the largest components, as the photograph clearly shows. Layout is not critical. Only the timing potentiometer (1 megohm) goes to the front panel; the other two are PC-board types. IC sockets are handy, and I even mounted the switching transistors in a single eight-pin socket.

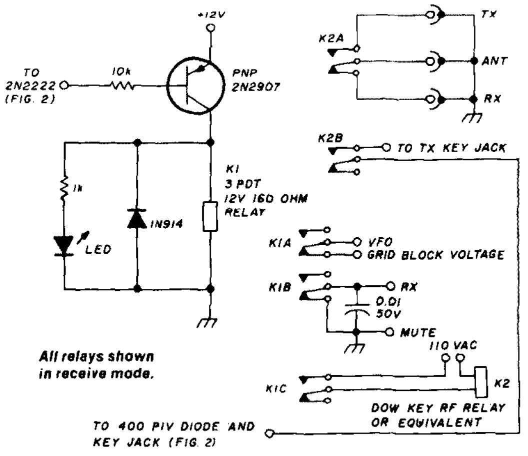


fig. 3. Modification of SSCWX for use of a Dow Key or similar rf relay.

The most critical components may be the 0.01-μF bypass capacitors. Voltage ratings for these capacitors are indicated, because some lines may carry fairly high voltages. This is true of the ac line and the key line at both its input and output points. All control lines running from the CWX relay are shielded and bypassed. The object is to be as certain as possible that rf cannot enter the CWX and disrupt any of the circuits.

The power supply diodes are 1N4002 100-PIV units. 1N914s work well in the timing circuit. The key line requires diodes with a PIV rating twice that of the anticipated transmitter key line voltage. 1N4004 (400-PIV) is specified, but anything with a higher voltage rating will work as well. Finally, be sure not to omit the reverse diode across the relay coil; it prevents reverse voltages generated by the collapsing relay coil field from reaching the switching transistor.

Besides the timing potentiometer, the front panel contains only the ac switch, the key jack, and two LEDs, one to indicate that power is on, the other to show that the relay is closed. Although it is possible to use a timing pot with an ac switch, the separate power switch permits you to retain your timing setting from one day to the next. In addition, you might want to consider adding a SPST toggle switch from the + 12 volt line to the relay in the collector circuit of the 2N2907; this would provide manual relay switching. The photo shows a phono jack used for the key, but any type will do.

The perf-board construction technique used for this project is good for one-of-a-kind projects in which lead length is no problem. As a group or club project, though, the circuit lends itself to printed-circuit-board fabrication. Build the circuit one step at a time, starting with the power supplies. If you build the remaining circuitry as a unit, at the very least test it progressively by plugging in only one solid-state device between checks. The 1k resistor in the TIL-116 diode line should work for most opto-isolators. The desired value is the highest that will just move the collector voltage (pin 5) from full to zero to full again under keying.

Next, add the LM311 and set the voltage comparator's minus input voltage to about 7 volts. Check the maximum time delay by measuring the output voltage swing. If the time is too long for your taste, increase the negative line reference voltage; if too short, decrease it. Perform these tests by holding the key down for at least one second to ensure the timing capacitor is fully charged, then release and time the delay. Be sure also that the supplementary charging voltage is at least a volt below the reference voltage. Once the long time delay is set, you can set the 10k pot so that the supplementary charging voltage

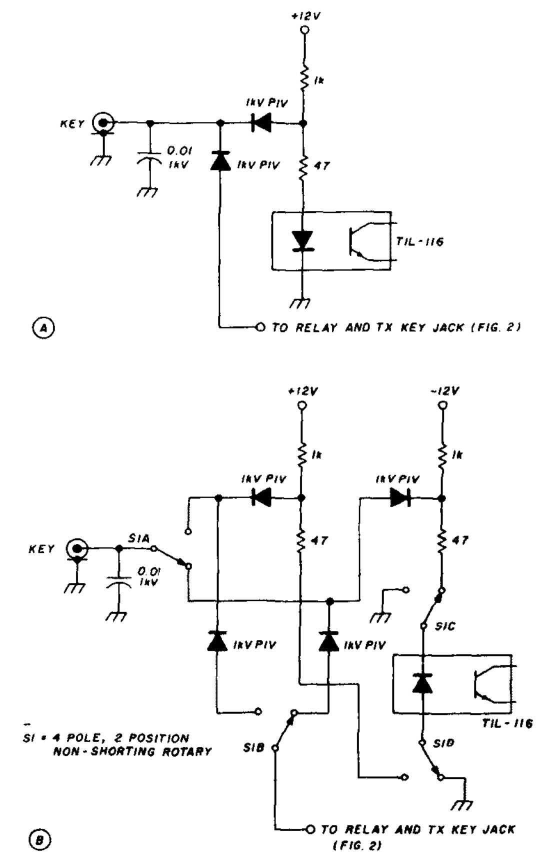


fig. 4. Modification of SSCWX key input. See text.

is between a half and a full volt less than the reference voltage.

Now check the delay at the minimum end of the line. If the timing pot allows the relay to open and close with every dit for more than 10 degrees of pot adjustment, then increase the series fixed resistor above the 39k value shown. If the relay does not come close enough to following your keying at minimum pot setting, then decrease the value of the fixed resistor. Finally, recheck the timing at the maximum end of the scale and do any final tweaking necessary.

Add the switching transistors, one at a time, and check their operation by measuring the collector voltages. When all is well, plug in the relay. If it operates well, mount the unit in the case. In the model shown in the photo, I set the perf board on 1-inch threaded pillars to keep the relay socket contacts well clear of the case. The only solder connection needed was for the key jack, since all other leads, indicators, and controls pass through the front panel from the rear.

The final test is with your transmitter and receiver.

Plug your key into the SSCWX unit. Be sure that the transmitter key jack is empty initially. Check the receiver mute line to be certain that the relay quiets the unit with the key down. Next, check the antenna relay for correct operation. The VFO control line is next. Finally, with a dummy load, plug the key line into the transmitter key jack. Only after you have tested the unit in this sequence should you put it on the air. In fact, you might want to unplug the key line from the transmitter (or shut the transmitter off) and get the feel of operating with the SSCWX, using only your receiver. It takes a little while to overcome the urge to grab a switch when going from transmit to receive and back again.

modifications and other uses

Listed below are some optional modifications that make the basic circuit a more versatile tool around the shack, and applicable to many projects.

Fig. 4A shows a revision to the input circuit for positive keying voltages. Simply reverse the leads to the diode section of the TIL-116 and change the supply voltage to +12 volts. As shown in **fig. 4B**, a 4PDT non-shorting rotary switch permits easy conversion from negative to positive line keying. You may want to add indicator lights to this circuit.

The basic circuitry of SSCWX lends itself to almost any timed relay control application. Where isolation is unnecessary, a common NPN switching transistor can replace the opto-isolator. At the other end of the line, small relays with light coil-current demands do not require a double transistor switch. Placing the relay in the collector circuit of the 2N2222 provides reliable service for up to about 50 mA of coil current. This arrangement provides excellent relay switching for CMOS controlled circuits, and the scheme is easily converted to TTL 5-volt levels.

To provide manual relay starting with automatic dropout, we can add a small SCR to the circuit, as shown in **fig. 5**. Any small plastic (TO-92) or metal (TO-5 to TO-18) case SCR, such as the 2N877 (rated at 30 volts and 0.5 amp) will work well in the circuit;

part	value or specification	Radio Shack No.
resistors	1 22-ohm, 1/4 watt	
	1 47-ohm, 1/4 watt	
	3 1-kilohm, 1/4 watt	
	2 2.2-kilohm, 1/4 watt	
	1 4.7-kilohm, 1/4 watt	
	4 10-kilohm, 1/4 watt	
	1 30-kilohm, 1/4 watt	
	1 39-kilohm, 1/4 watt	
	1 47-kilohm, 1/4 watt	
potentiometers	1 1-megohm, panel mounting type	271-211
	1 10-kilohm, pc board type	271-335
	1 100-kilohm, pc board type	271-338
capacitors	2 1000-μF, 25-volt electrolytic	272-1019
	1 1-μF, 16-volt tantalum	
	1 5-μF, 16-volt electrolytic	272-1024
	4 0.01-μF, 1 kV disc	
	2 0.01-μF, 50-volt disc	
diodes	2 400-PIV (1N4004)	276-1103
	2 100-PIV (1N4002)	276-1101
	3 1N914 switching diodes, silicon	278-1122
	1 1N4742, 12-volt, 1-watt Zener	276-563
ICs	1 7812 positive 12-volt regulator	276-1771
	1 LM311	
	1 TIL-116 or TIL-111	276-132
transistors	1 2N2222 NPN, 40-volt, 100 mA	276-1617
	1 2N2907 PNP, 40-volt, 500-mA	
relay	1 4PDT 12-volt, 160-ohms	275-214
	1 socket for above	275-221
	1 DPDT 12-volt, 160-ohms, 10 amp contacts	275-218
transformer	1 12.8-volt, 300-mA secondary	273-1305
switch	1 SPST miniature toggle	275-624
fuse	1 1/8 amp, slow blow	
LED	2 0.2-inch-diameter, with panel lens	
miscellaneous	3 8-pin DIP IC sockets	
	hardware	
	perf board	
	case (5-1/4 × 3 × 5-7/8 inches)	270-253
	key jack and output connectors or cables to suit builder	

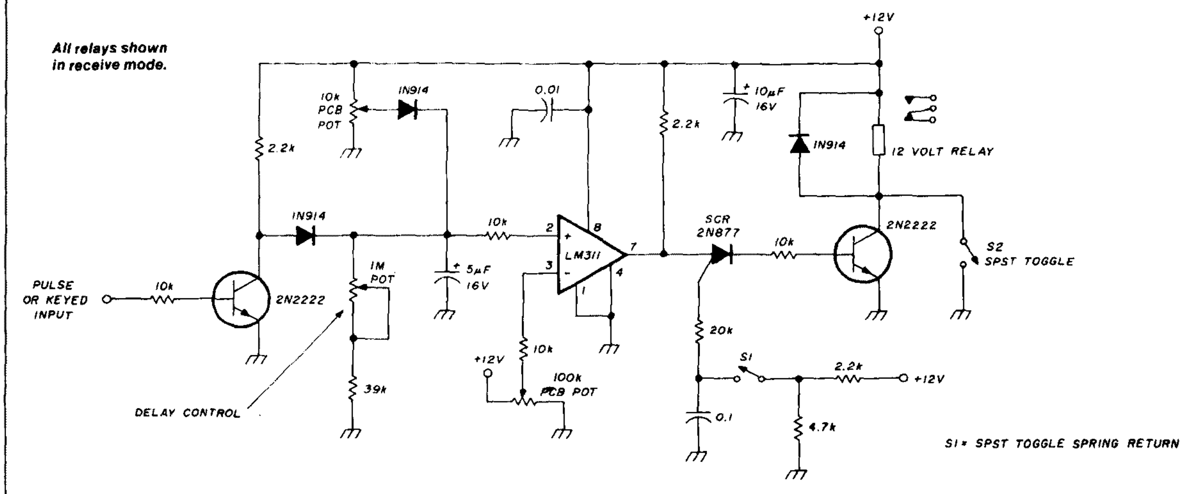


fig. 5. Schematic diagram of a relay switching circuit with SCR control.

avoid high-power SCRs. S1, a spring-return SPST toggle switch, provides triggering voltage to the SCR, which then conducts until the output of the LM311 goes low (even though the switch has been released). In short, the SCR blocks the comparator output from the relay transistor until fired by a manual touch of the switch, but then the SCR continues conducting until the input signal disappears and the time delay set by the LM311 input circuitry ends. A second SPST toggle, S2, provides manual relay control.

The SCR gate circuitry provides a low voltage and current to trigger the gate. For more rapid gate voltage decay after release of S1, parallel a 100 kilohm, 1/4-watt resistor across the 0.1-μF capacitor. Although all circuit values have worked well for several unmarked SCRs in TO-92 cases, adjust the voltage divider and series gate resistor values to suit the devices you have on hand.

The circuit in **fig. 5** has numerous uses. Without the SCR, a train of digital pulses will key and hold the relay until a specified time after the last one. With the SCR, the relay pulls in upon manual starting with S1, a handy feature with very narrow passband CW audio filters. The use of an LED indicator in the filter, to indicate visually when the signal is locked, permits switch-over to the filter with no loss of copy while trying to tune the filter to the signal (or vice versa). A similar system can eliminate readout garbage in RTTY systems.

The SSCWX switching system provides a sound basis for designing time-controlled relay switching systems for many applications. Apart from the relay, which may range in size from DIP scale to one inch square and two high, the circuit requires a couple of square inches of perf or etched board space, plus the panel pot and any toggle switches. The solid state CWX unit is small enough (including power supply) to mount inside the cabinet of most older transmitters and receivers. There is room on most front panels for the few controls required. On the DX-60, for example, the audio gain control is convertible to CWX duty if the a-m feature is dispensable. A little rewiring of the key jack finishes the job, since the indicator lights are no longer needed. The only precaution is to be sure that the SSCWX is well shielded and bypassed from the DX-60's rf.

Hidden or in the open, the SSCWX makes a good addition to any shack using separate units to transmit and receive. The ease my old CWX gave to my CW operations back in the '60s made it deserving of an update. SSCWX is that update. The basic circuit and its variations can solve relay switching problems for most any ham or experimenter.

ham radio

capacitively coupled hybrids

These devices can be used to divide, combine, or phase-shift power for Amateur applications

Do you need a device that can split or combine power? How about a phase shifter or a circular polarizer? Power can be divided, combined, sampled, or phase-shifted in any increment by using a hybrid. A capacitively coupled hybrid is easily constructed and very compact.

The capacitively coupled hybrid is a four-port device consisting of two transmission lines. Output power at one port of the coupled transmission line is dependent on the direction of propagation in the main transmission line. In **fig. 1**, a wave traveling from port 1 (input) to port 2 is coupled to port 4. Ideally, no power appears at the isolation port 3. A wave traveling in the opposite direction, from port 2 to port 1, is coupled to port 3, indicating any mismatch at port 4. The coupling factor determines the amount of power coupled from the main line to the power output at the coupled-line output port, assuming matched loads present at all ports.

General expression

$$Coupling = 10 \log_{10} (P_1/P_4) dB \qquad (1)$$

Coupling factor for the capacitively coupled hybrid

$$Coupling\ (dB) = 20 \log_{10} \cos\theta \qquad (2)$$

where θ is the electrical length of the transmission line. The capacitive reactance of the coupling capacitor is:

$$X_c = Z_o \cdot \tan\theta \qquad (3)$$

where Z_o is the impedance of the transmission line. The line impedance is equal to the termination impedance at each port. The value of capacitance is:

$$C_c = \frac{1}{2\pi f X_c} = \frac{1}{2\pi f Z_o \tan\theta} \qquad (4)$$

Because the device is reciprocal, a wave incident at port 2 is coupled to port 3 by the same coupling factor.

The hybrid described here is smaller than the Wilkinson hybrid (described in an earlier article[1]), easy to construct, and displays improved amplifier performance. The principal disadvantage of this hybrid is its limited bandwidth of 10 percent. This, however, is still adequate for most Amateur use.

coupler operation

A wave traveling to the right (port 1 to port 2) on the main transmission line will have a portion of its energy coupled to the second transmission line at each end by capacitors. It is assumed that there is no inductive coupling. The two signals arriving from both paths at port 4 are in phase and combine. The signals at port 3 are also equal (**fig. 1B**) but 180 degrees out-of-phase and cancel. The longer path always has 180 degrees more phase shift because of the transmission lines. (The phase shifts that result

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from the coupling capacitors cancel out because they are present in both paths.)

In case of a wave traveling in the reverse direction (port 2 to port 1), the coupled energy combines at port 3 and cancels at port 4, with the capacitively coupled transmission lines acting as a directional coupler. A sample of the energy from a forward wave appears at port 4. A sample of the reflected wave appears at port 3. The ratio of power at port 3 to power at port 4 yields the return loss.

$$return\ loss = -10 \log_{10} (P3/P4)$$

$$= -10 \log_{10} (reflected\ power/forward\ power)$$

directional coupler

One's first thought would probably be to use the hybrid for measuring standing wave ratios. A detector placed at port 4 would indicate forward power, while a detector at port 3 would indicate reflected power. But how isolated are the two readings? How well is port 3 isolated from port 4?

Directivity is a measure of how well port 3 is isolated from port 4 with respect to power entering at terminal 1. Directivity is defined as:

$$directivity = 10 \log_{10} \frac{P_4}{P_3} \qquad (6)$$

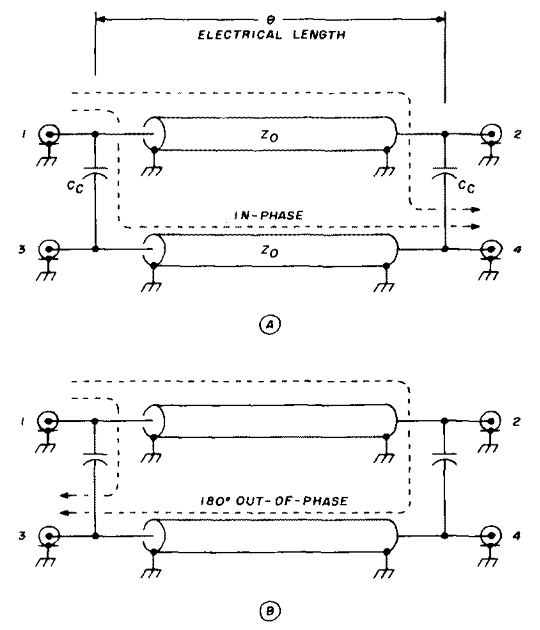


fig. 1. The capacitively coupled hybrid functions because of phased combining of signals from different paths. See text.

Ideally, the directivity should be infinite — but a directional coupler with a directivity of 20 to 30 dB is quite good. How high a directivity can be achieved is strongly dependent on the match at port 4, since reflected power is readily transmitted back to port 3.

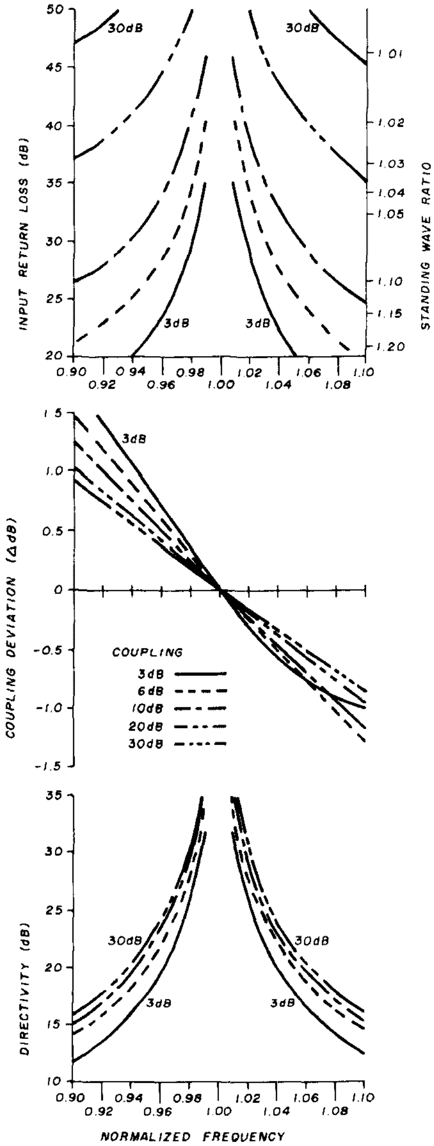


fig. 2. The theoretical response of the hybrid shows that, for reasonable performance, it has about a 10 percent bandwidth.

The theoretical response of the capacitively coupled hybrid is shown in **fig. 2**. The input return loss and directivity are better than 20 dB (VSWR ≤ 1.2:1) over a 10 percent bandwidth. The deviation from its nominal coupling is plotted for coupling from port 1

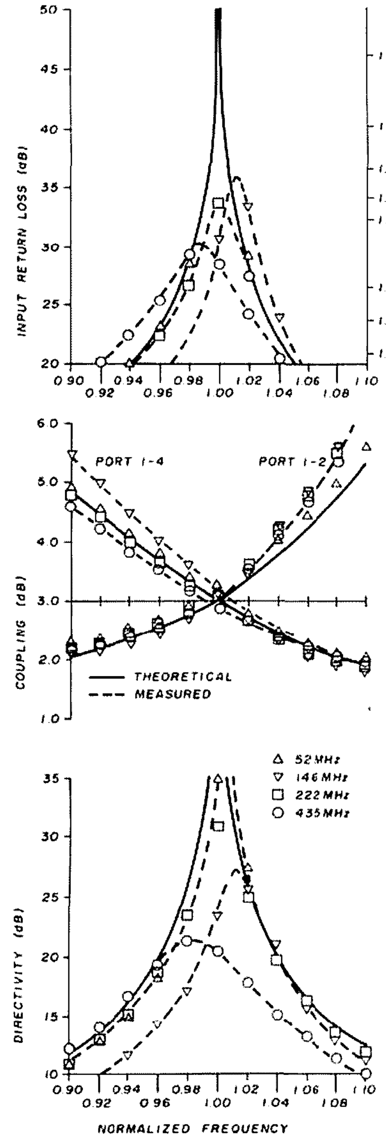
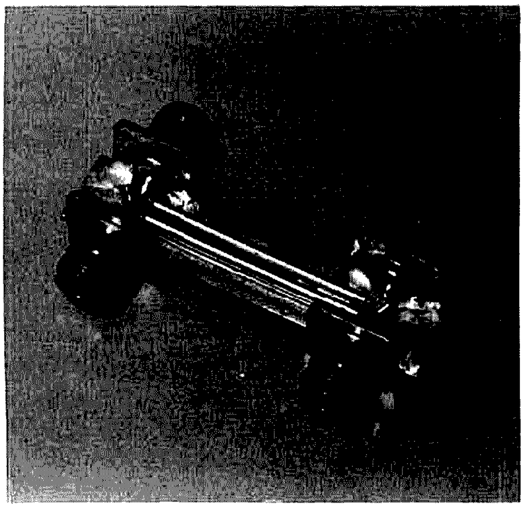
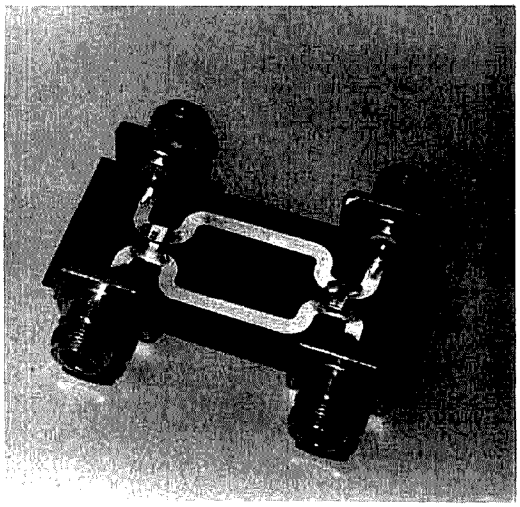


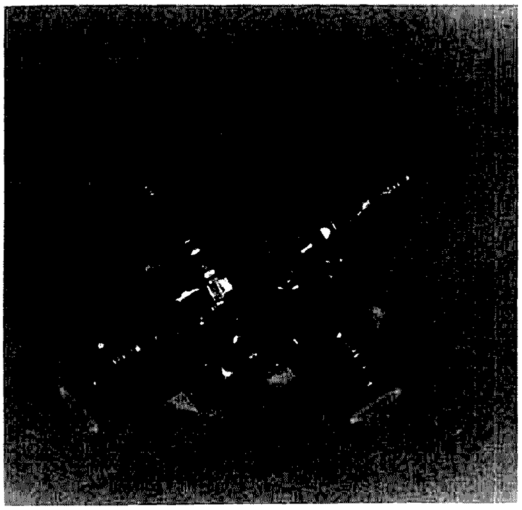
fig. 3. Models constructed using semi-rigid coax perform close to theoretical at VHF.



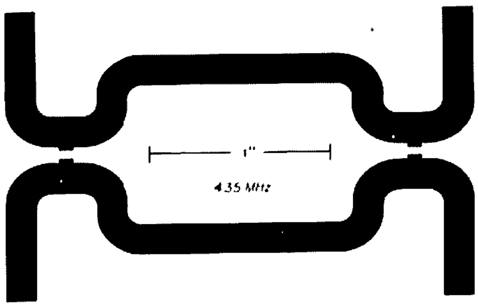
Micro strip hybrid.



Transmission line hybrid.



Lumped-constant hybrid.



Artwork for 435-MHz 3 dB coupler.

table 1. Design equations for transmission line and pi lumped-constant hybrids.

$$\theta = \cos^{-1} 10^{-K/20}$$

$$C_c = \frac{1}{2\pi f Z_o \tan\theta}$$

transmission-line hybrid

$$L = \left(\frac{Z_o}{2\pi f}\right) \sqrt{1 - 10^{-K/10}}$$

$$C_s = \left(\frac{1}{2\pi f Z_o}\right) \sqrt{\frac{10^{K/20} - 1}{10^{K/20} + 1}}$$

$$C_c = \left(\frac{1}{2\pi f Z_o}\right) \sqrt{\frac{1}{10^{K/10} - 1}}$$

pi lumped-constant hybrid

where K = $10 \log_{10} (P_1/P_4)$ absolute coupling
Z_o = termination impedances
θ = electrical transmission line length

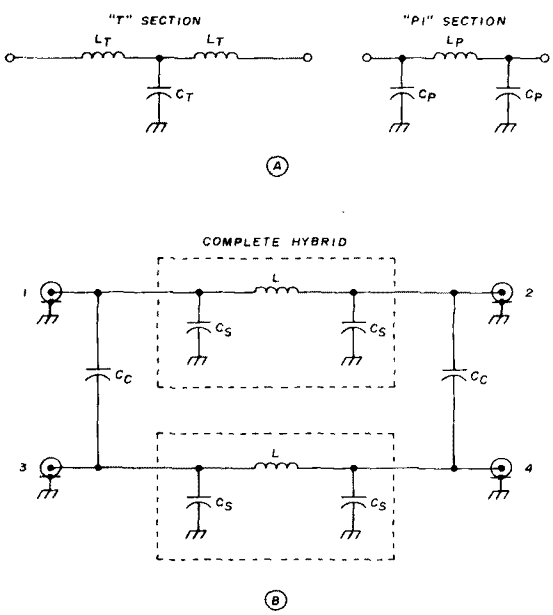


fig. 4. (A) The hybrid may be formed by using the T or pi section artificial lines. (B) A complete hybrid.

to port 4. The results from several experimental 3 dB models formed using eighth-wave semi-rigid transmission lines are shown in **fig. 3**. The theoretical response is shown as a solid line. Photographs are shown of transmission line, lumped-constant, and

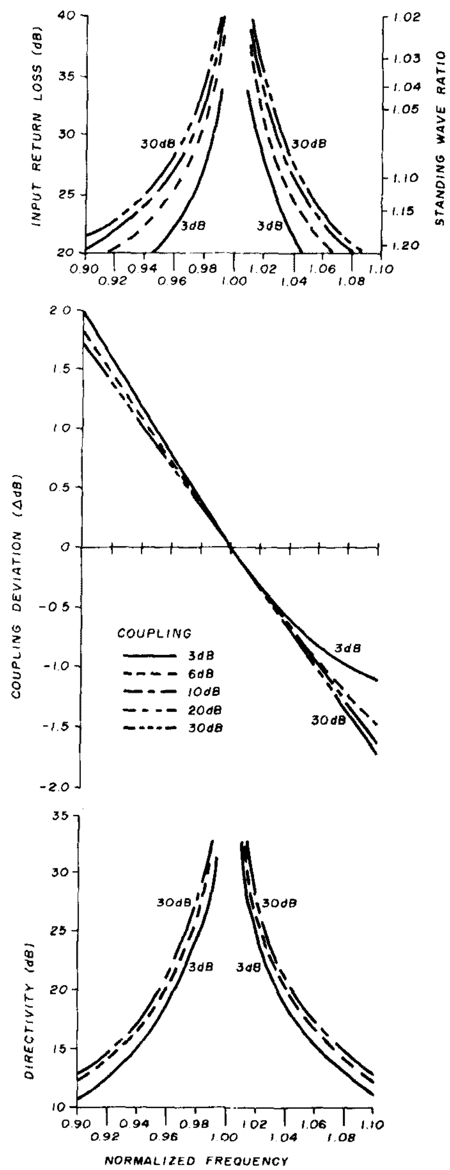


fig. 5. The theoretical response of the lumped-constant version of the hybrid is not as good as that of the transmission-line model.

microstrip hybrids. The artwork for a 435-MHz, 3-dB coupler is also shown for use on 1/16-inch, one-ounce Teflon-glass printed-wire board.

lumped constant coupler

The capacitively coupled hybrid may be formed with lumped constants in place of the coaxial cables. An artificial transmission line may be formed using lumped constants configured as a T section or a pi section, as shown in **fig. 4A**. The pi section is usually chosen to simulate the transmission line because it involves only one inductor and acts as a lowpass filter. The lowpass filter decreases harmonic energy, and inductors are more difficult to measure exactly than capacitors.

The values for the T artificial line are given as:

$$L_T = Z(1 - \cos\theta)/2\pi f \sin\theta \quad \textbf{(7)}$$

$$C_T = \sin\theta / Z\, 2\pi f \quad \textbf{(8)}$$

where Z refers to the input and output impedances and θ is the phase delay through the network.

For a pi network the needed inductance and capacitance are calculated from:

$$L_p = Z \sin\theta / 2\pi f \quad \textbf{(9)}$$

$$C_p = (1 - \cos\theta)/Z\, 2\pi f \sin\theta \quad \textbf{(10)}$$

table 2. Component values for transmission-line and pi lumped-constant hybrids.

coupling (dB)	line length θ (deg.)	fractional wavelength (λ/λ_o)	coupling capacitance (÷ f in MHz) (pF)	series inductance (÷ f in MHz) (nH)	shunt capacitance (÷ f in MHz) (pF)
3	45.0	0.125	3183	5,627	1,318
4	50.9	0.141	2589	6,174	1,514
5	55.8	0.155	2165	6,580	1,685
6	60.0	0.167	1838	6,886	1,835
10	71.6	0.199	1061	7,549	2,294
15	79.8	0.222	575.2	7,831	2,659
20	84.3	0.234	320	7,918	2,879
25	86.8	0.241	179.3	7,945	3,007
30	88.2	0.245	100.7	7,954	3,084

table 3. Component values for Amateur band transmission-line and pi lumped-constant hybrids.

frequency (MHz)	coupling (dB)	coupling capacitance (pF)	series inductance (nH)	shunt capacitance (pF)
3.750	3	849	1,500	352
	20	85.3	2,112	768
7.150	3	445	787	184
	20	44.7	1,107	403
14.175	3	225	397	930
	20	22.6	559	203
21.225	3	150	265	62.1
	20	15.1	373	136
28.850	3	110	195	45.7
	20	11.1	274	99.8
52	3	61.2	108	25.4
	20	6.15	152	55.4
146	3	21.80	38.50	9.03
	20	2.19	54.20	19.7
222	3	14.3	25.30	5.94
	20	1.44	35.70	13
435	3	7.37	12.90	3.03
	20	0.74	18.20	6.62
1296	3	2.46	4.34	1.02
	20	0.25	6.11	2.22

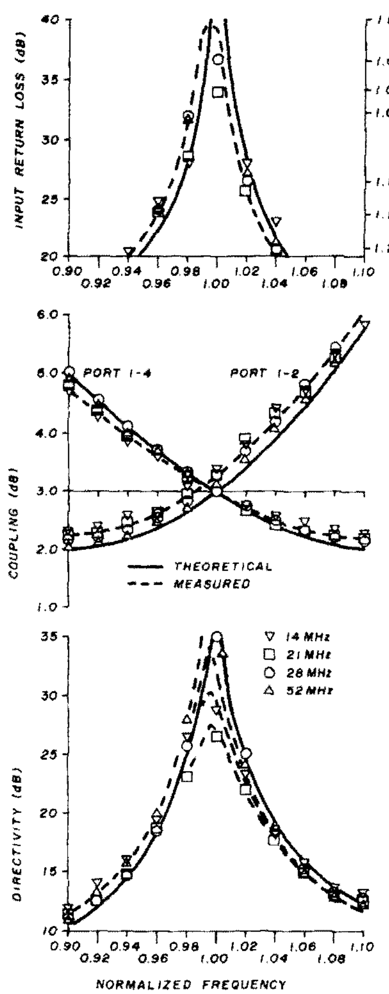


fig. 6. Several equal-split hybrids were constructed using lumped constants.

By substituting the coupling equation,

$$\cos\theta = 10^{-K/20} \tag{11}$$

where K is the coupling between port 1 and port 4 in decibels, into the above equations for the pi network and using a few trigonometry identities, we arrive at the design equations for the lumped-constant, capacitively coupled hybrid given in **table 1**.

A capacitively coupled hybrid is shown in **fig. 4B**. The predicted response of the lumped constant version is shown in **fig. 5**. The graph of the input return loss shows that the lumped constant version has slightly less bandwidth than the lightly coupled transmission line version ($\geq$ 10 dB). At frequencies below 100 MHz the length of transmission line is too great for compact implementation. Values for the construction of transmission-line or lumped-constant, capacitively coupled hybrids are catalogued in **table 2**. Values for each of the Amateur bands are given in **table 3**.

Lumped-constant models were constructed for several Amateur bands. The results for 3-dB and 20-dB couplers are shown in **figs. 6 and 7**. The 3-dB hybrid is commonly used for power splitting/combining, while the 20-dB hybrid is normally used as a

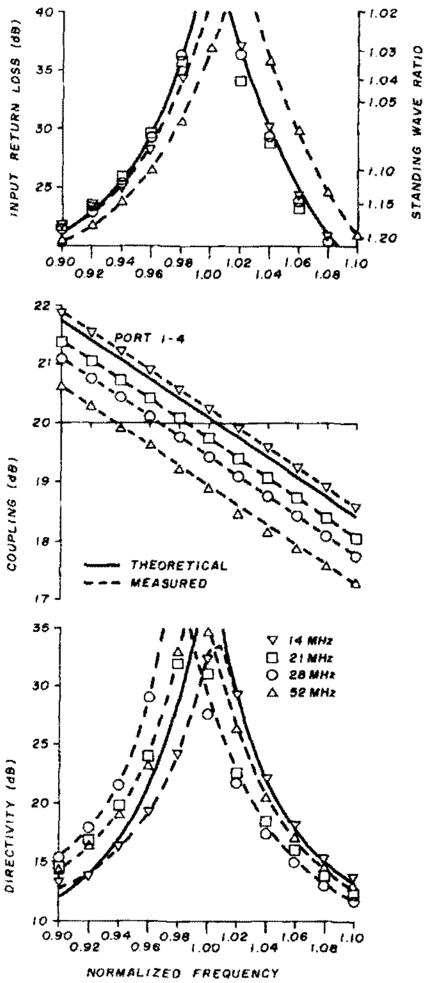


fig. 7. The 20-dB hybrid is used mainly for sampling power in a transmission line.

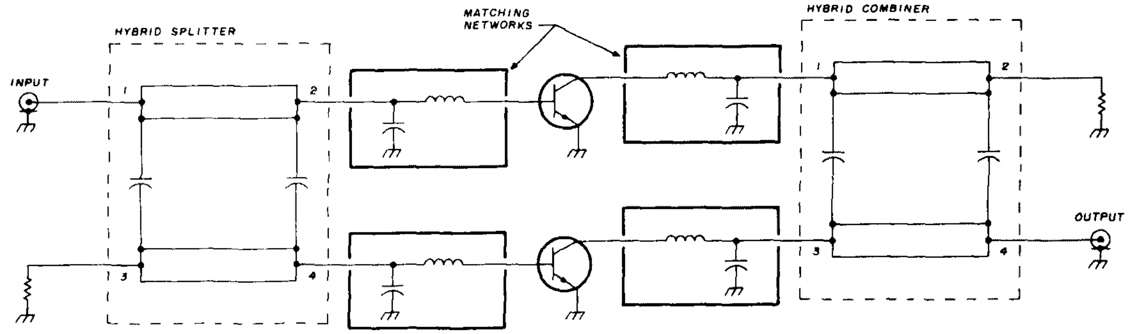


fig. 8. The most common use for a hybrid is splitting or combining power, such as paralleling two transistor amplifiers.

directional coupler. The results agree quite well with the theoretical values shown as a solid line. The responses could be adjusted for perfect agreement if one used greater precision in component selection. Capacitors with marked values close to the calculated values were used in these hybrids. With matched terminations on each port, the inductor closest to the input port is adjusted for best return loss. The input was then applied to port 4 with all the other ports terminated and the match adjusted using the other inductor. Spreading or squeezing of the inductor is determined using a "diddle" stick. This stick is constructed by gluing a piece of brass to one end of a plastic stick and a piece of ferrite to the other end. If the performance improves with the proximity of brass, the coil needs to be spread for less inductance. The coupling between transmission lines is mainly a factor of the coupling capacitance. The capacitance is easily measured using a digital capacitance meter.

The most popular use for the capacitively coupled hybrid is splitting the input and combining the output power from two transistors in a power amplifier, see **fig. 8**. The transistors remain isolated from each other. Isolation is a measure of how much of the power incident at port 1 appears at the isolation port 3. It is also the same as the reflected power at port 4 appearing at port 2, because the device is reciprocal. Isolation is equal to the coupling (port 1 to port 4) plus the directivity. For this hybrid the isolation between ports 1 and 3 or between ports 2 and 4 is the same as the input return loss. Thus one can expect the mismatched power at the input to one transistor to be reduced by 20 dB (X100) before appearing at the input to the other transistor. This isolation improves stability by decreasing interaction. Any mismatch difference at the input to the two transistors appears at the isolation port. If the mismatch is identical the hybrid still provides a good input match, and no power appears at the isolator port. Power at the isolation port of the output hybrid indicates gain differences of the transistors.

intermodulation performance

With respect to a signal applied to input port 1, the output signals at ports 2 and 4 on a 3-dB hybrid are 90 degrees out of phase. One advantage that the 90-degree hybrid has over the Wilkinson combiner is the improvement in amplifier output intermodulation performance. A signal (F_1) from a nearby interfering transmitter coupled through the transmitting antennas mixes with the second harmonic of the amplifier ($2F_0$) at the collector in a power amplifier to form an intermodulation product ($2F_0 \pm F_1$). This mixing product is typically close to the desired carrier frequency and thus very difficult to filter. Examination of the 90-degree hybrid shows that the intermodulation products (F_{IM}) from the amplifier shown

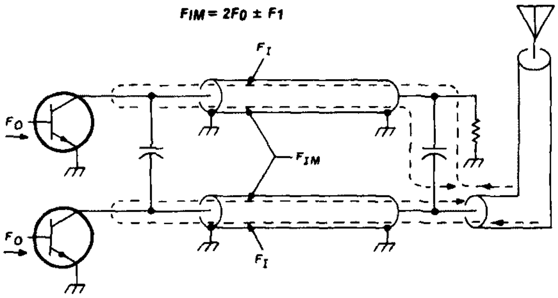


fig. 9. Transmitter intermodulation is reduced because the intermodulation products combine 180 degrees out of phase at the output port.

in **fig. 8** combine at the output port 180 degrees out of phase to cancel out, **fig. 9**. The round trip path of one interference-intermodulation signal is 180 degrees longer than the round trip path to the other transistor.

if peak detector diodes are used. A 3-dB quadrature hybrid may be used to induce circular polarization[3] in a pair of crossed dipoles, **fig. 10C**. By using a 3-dB hybrid to establish a 90-degree phase shift, a receiver mixer[4] can be formed, **fig. 10D**. The input signal and local oscillator are isolated typically by 30 dB over a 4 percent bandwidth. The additional quarter-wave line at port 2 is used to establish an additional 90-degree phase shift with respect to port 4. Thus the mixer diodes are driven 180 degrees out of phase with respect to each other and the local oscillator signal is balanced at the i-f port.

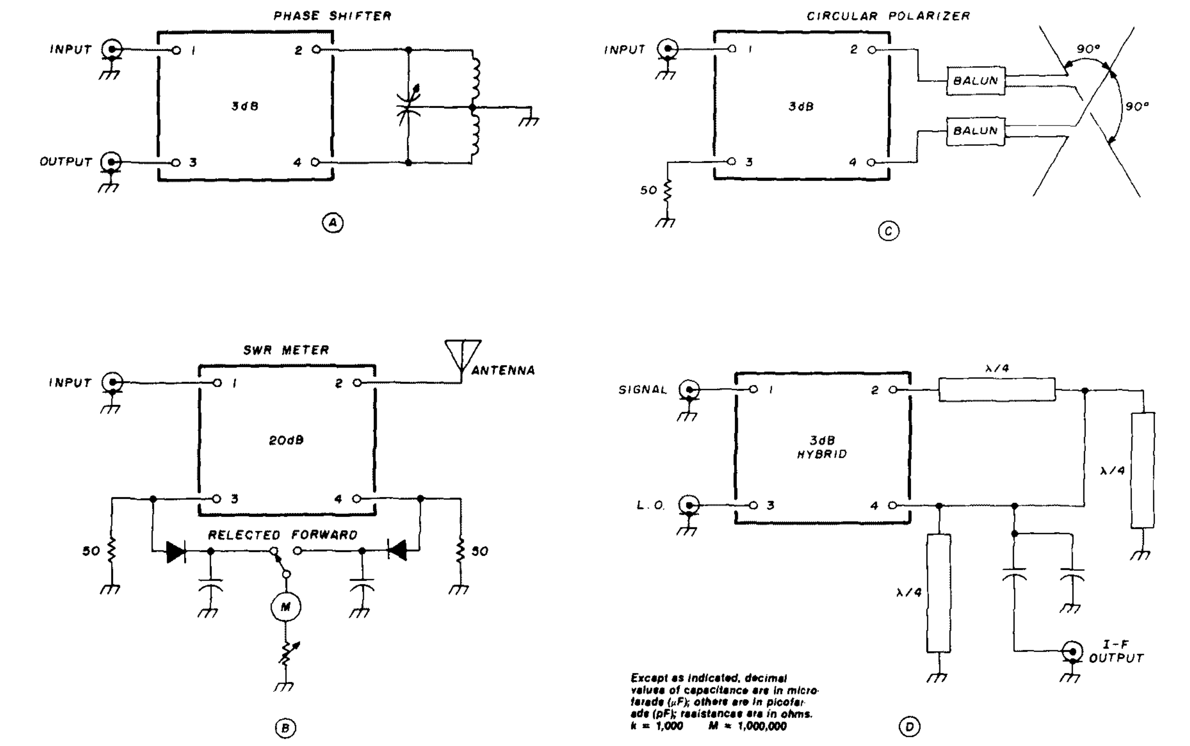


fig. 10. The capacitively coupled hybrid may be used as an (A) phase shifter, (B) SWR meter, (C) circular polarizer, and (D) mixer.

As long as each collector provides identical terminations to the combiner hybrid, the level of the intermodulation signal at the output port should be reduced by the value of the input return loss, graphed for each hybrid.

other uses

The 3-dB coupler may be used as a phase shifter for varying the phase in one leg of a phased array.[2] See **fig. 10A**. These antenna systems typically cover a single Amateur band. By varying the purely reactive load at ports 2 and 4 the phase may be adjusted over 180 degrees. The standing wave ratio coupler shown in **fig. 10B** uses a 20-dB hybrid. The measured return loss is given as

$$\text{return loss (dB)} = -20 \log_{10} \frac{V_3}{V_4} \qquad (12)$$

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2. Henry S. Keen, W5TRS, "Electrically Controlled Phased Array," *ham radio*, May, 1975, pages 52-55.
3. Reed E. Fisher, W2CQH, and Richard H. Turrin, W2IMU, "UHF Directional Couplers," *QST*, September, 1970, pages 26-31.
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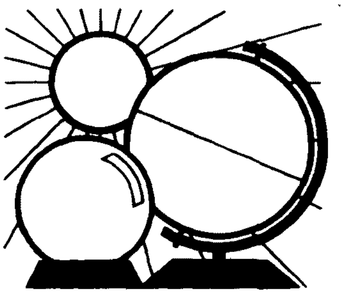

DX FORECASTER

Garth Stonehocker, KØRYW

last minute forecast

DX conditions on 10, 15, and 20 meters are expected to peak during the second and third weeks of this month because of high MUFs resulting from increased solar activity. Disturbed conditions, because of solar flare emissions, are expected on the 10th and 18th. Other periods of geomagnetic and ionospheric (recurrent) disturbances are likely the 1st and 27th because of enhanced solar wind from coronal holes. Nighttime DX on the lower bands should continue to be good during the entire month, with best performance during the first and last weeks.

With the advent of the vernal equinox this month (21st), gray-line DXing should occur during both local sunrise and sunset. North/south paths over the polar regions should be useful, during lulls in geomagnetic storm activity. Geomagnetic disturbances, which are more evident as the equinox approaches, cause considerable signal attenuation and fading on polar paths. (Gray-line operation is explained in *ham radio*, September, 1982, page 56, and *CQ Magazine*, September, 1975, page 27).

band-by-band summary

Ten, fifteen, and twenty meters will be open from morning to early evening almost every day, and to most areas of the world. The openings will be shorter on the higher bands and occur more frequently at local noon. Trans-equatorial propagation will be more likely on these bands during conditions of high solar flux and a disturbed geomagnetic field.

Thirty meters will be useful almost twenty-four hours a day. Daytime conditions will resemble those on 20 meters, except that signal strengths may decrease during midday on some days, those days coinciding with high solar flux values. Nighttime use will be good except following days of very high MUF conditions. Generally, the usable distance is expected to be greater than achieved on 80 at night but less than that on 20 meters during the day.

Forty, eighty, and one-sixty meters are the night DXer's bands. The bands are open just before sunset and last until sunrise, local time. Except for daytime short-skip signal strengths, high solar flux values won't greatly affect these bands.

MARCH

WESTERN USA

GMT	PST	N (Asia, Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania, Australia)	NW (Japan)
0000	4:00	10	20	15	10	15	10	10	10
0100	5:00	10	20	15	10	15	10	10	10
0200	6:00	15	40	20	10	15	10	10	10
0300	7:00	15	40	20	15	15	10	10	15
0400	8:00	15	40	20	15	15	10	10	15
0500	9:00	20	40	20	20	15	10	15	20
0600	10:00	—	40	—	20	15	15	15	20
0700	11:00	—	—	—	20	20	15	20	40
0800	12:00	—	—	20	20	20	20	20	40
0900	1:00	—	—	20	20	20	20	20	40
1000	2:00	—	—	—	20	40	20	20	40
1100	3:00	—	—	—	—	40	20	20	—
1200	4:00	—	—	—	—	40	40	20	—
1300	5:00	—	—	—	—	—	40	20	—
1400	6:00	—	20	20	15	—	—	—	—
1500	7:00	—	15	15	10	—	—	—	—
1600	8:00	20	15	10	10	—	—	15	—
1700	9:00	20	15	10	10	—	15	15	—
1800	10:00	20	15	10	10	—	10	15	—
1900	11:00	15	15	10	10	—	10	15	—
2000	12:00	15	20	10	10	15	10	15	20
2100	1:00	15	20	15	10	15	10	10	15
2200	2:00	10	20	15	10	15	10	10	10
2300	3:00	10	20	15	10	15	10	10	10

MID USA

GMT	MST	N (Asia, Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania, Australia)	NW (Japan)	CST
0000	5:00	15	20	15	10	15*	10	10	10	6:00
0100	6:00	15	20	15	10	15	10	10	10	7:00
0200	7:00	15	40	20*	15	15	10	10	15*	8:00
0300	8:00	20	40	20	15	15	10	15	15	9:00
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0700	12:00	—	20	20	20	20	20	20	40	1:00
0800	1:00	—	—	20	20	20	20	20	40	2:00
0900	2:00	—	—	—	—	20	20	20	40	3:00
1000	3:00	—	—	—	—	40	40	—	—	4:00
1100	4:00	—	—	—	20	—	40	—	—	5:00
1200	5:00	—	—	15	15	—	—	—	—	6:00
1300	6:00	20	—	15	15	—	—	—	—	7:00
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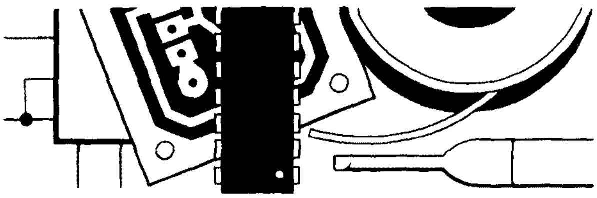
2000	1:00	—	15	10	10	15	10	10	20	2:00
2100	2:00	—	20	10	10	15	10	10	15	3:00
2200	3:00	—	20	10	10	15	10	10	10	4:00
2300	4:00	—	20	10	10	15*	10	10	10	5:00

EASTERN USA

GMT	EST	N (Asia, Far East)	NE (Europe)	E (S. Africa)	SE (Caribbean, S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania, Australia)	NW (Japan)
0000	7:00	15	20	15	10	15	10	10	15
0100	8:00	15	20	15	10	15	10	10	15
0200	9:00	15	40	20	15	15	10	15	15
0300	10:00	15	40	20	15	15	15	15	20
0400	11:00	20	40	15	15	20	15	20	20
0500	12:00	—	40	15	20	20	20	20	20
0600	1:00	—	20	20	20	20	20	20	40
0700	2:00	—	20	—	20	20	20	20	40
0800	3:00	—	20	—	20	20	40	—	40
0900	4:00	—	—	—	20	20	40	—	—
1000	5:00	—	—	—	20	—	—	—	—
1100	6:00	—	—	15	15	—	—	—	—
1200	7:00	—	—	10	15*	—	—	20	—
1300	8:00	20	15	10	10	—	—	15	—
1400	9:00	20	10	10	10	—	—	15	—
1500	10:00	20	10	10	10	—	—	15	—
1600	11:00	15	10	10	10	—	—	15	—
1700	12:00	15	10	10	10	—	20	15	—
1800	1:00	15	15	10	10	—	20	15	—
1900	2:00	—	15	10	10	15	15	10	—
2000	3:00	—	15	10	10	15*	10	10	20
2100	4:00	—	15	10	10	10	10	10	15
2200	5:00	—	20	10	10	10	10	10	10
2300	6:00	15	20	10	10	10	10	10	10

*Look at next **higher** band for possible openings.

the weekender



CW zero-beat indicator for transceivers

For many CW operators who use transceivers, zero beating the station they are working is very difficult. I, too, had this problem, and after several years of searching for an answer I devised the circuit illustrated in **fig. 1**. It puts me within 50 Hz of the frequency of the station I am working.

Terminal A (see **fig. 1**) is connected to the audio output line of my transceiver. When I am zero beat with the CW station I'm working, the LED lights when the other operator keys his transmitter. If the LED does not light, my rig is not zero beat.

The LM567 (U1) contains an oscillator, with output frequency controlled by the value of R1 and C1. The value of R1 is adjusted so the LM567 oscillator frequency is equal to the offset frequency of your transceiver. The LM567 output is brought out at pin 8 and is an open collector. As you examine **fig. 1**, you will see that the collector load for the output transistor is the LED and R3, which are connected to +5 volts. When the input frequency to terminal A is equal to the LM567 oscillator frequency, the transistor is turned on and the LED lights up to provide an indication that the audio frequency entering terminal A is equal to the oscillator frequency.

The amplitude of the audio signal fed to terminal A will effect the bandwidth (the band of frequencies that will cause the LED to light). If the LM567 oscillator is one kHz, and the input signal at pin 3 is 300 $mV_{p\text{-}p}$, the bandwidth of this circuit will be approximately 350 Hz. Under this condition, you will get a zero beat indication when the audio input signal is at any frequency betwen 825 and 1175 Hz. So, you may be off-frequency by 175 Hz and still get a zero beat indication. We can do much better than this, however. A 25 $mV_{p\text{-}p}$ signal on pin 3 will provide a 100 Hz bandwidth and the most that you will be off is 50 Hz.

The lower the amplitude of the audio signal arriving at pin 3 the more accurate you will be in zero beating the station you are working. However, you will still want to hear the audio signal, so R2, CR1 and CR2 function to limit the maximum amplitude of the audio signal at pin 3 of the LM567 to a usable level while maintaining adequate volume from the speaker or headphones. My HW-101 has an 8-ohm audio output, and I find the 100-ohm resistor value for R2 adequate. However, 100 ohms may be too high for use with a 4-ohm audio output stage. Therefore, it may be desirable to install a 500-ohm potentiometer, and set the resistance of R2 to suit your own taste.

The capacitance of C1 must be temperature stable. I had the sad experience of using a capacitor for C1 with capacitance that changed with temperature, and the oscillator frequency of the LM567 changed

By Glen Carlson, W6KVD, 2588 Hermosa Street, Pinole, California 94564

with the temperature of my rig. The capacitor type listed for C1 in the parts list works well for me.

Owners of solid-state rigs having a +5 Vdc supply will need only the circuit illustrated in fig. 1. Owners of tube type rigs may need to build a 5 volt supply, illustrated in fig. 2.

construction

I built the circuits illustrated in figs. 1 and 2 on a Radio Shack™ 276-024 printed circuit board. These boards are inexpensive and are made to accept a 16-pin IC. They provide pads at the edge of the board connected to the 16 IC pins. I recommend installing an 8-pin IC socket for IC1. Tin the traces on the board before drilling small holes in the pads.

I drilled a hole in the front panel of my HW-101 and installed a grommet to hold the LED in place. I used single-conductor shielded audio cable for running power to the LED. The shield must be isolated from the chassis as it will be used as one of the conductors to the LED.* Find a place for installing the PC-board, and then construct a small U-bracket for mounting the board while maintaining isolation from the chassis.

Tie terminal A to a point in your transceiver where this circuit will operate when using either the speaker or headphones. If you tie terminal A to a point where there is a dc voltage, CR1 or CR2 will be biased into conduction. In this case, use a suitable blocking capacitor between terminal A and the point you pick up the audio signal. A capacitor value between 0.01 μF and 0.1 μF should suffice. Adjust the value of R3 for suitable brilliance of the LED.

alignment

Alignment consists of setting the LM567 oscillator frequency equal to the offset frequency of your transceiver. There may be several ways to accomplish this adjustment. You will set the oscillator frequency by adjusting R1.

If you have a ham friend who can zero beat your transmit frequency, adjust R1 while listening to his signal. As you adjust R1, keep reducing the volume to narrow the bandpass for the best adjustment of the oscillator frequency. Then practice zero beating your friend's frequency, and have him check the accuracy of your adjustment.

A more accurate method uses a frequency counter. Measure the frequencies of the oscillators which determine the frequency difference between your transmit and receive frequencies. Subtract the smaller frequency from the larger frequency, and you will have your offset frequency. Using a low capacity probe, connect the frequency counter to pin 5 of the LM567, and adjust R1 until the counter reads your offset frequency.

When using the frequency counter method, you may be surprised to find that your offset frequency is not exactly as listed by the transceiver manufacturer. In my case, the offset frequency was off by 70 Hz.

If enough Amateurs build this circuit, we may find room in the CW bands for more QSOs.

ham radio

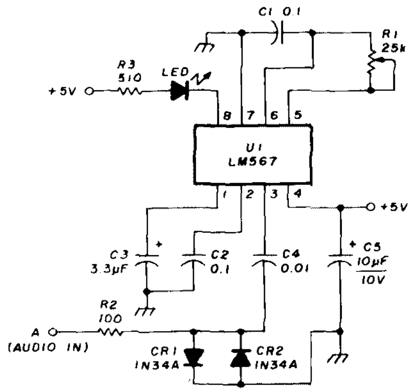


fig. 1. Terminal A, connected to audio output line.

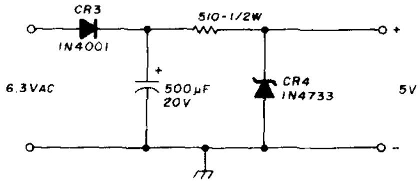


fig. 2. Five-volt power supply.

C1	0.1 μF 200 V (see text) RS 272-1053
C2	0.1 μF 10 V
C3	3.3 μF 10 V tantalum
C4	0.01 μF 10 V
C5	10 μF 10 V (tantalum)
C6	500 μF 20 V
CR1 & CR2	small signal germanium diodes
CR3	rectifier diode 1N4001 RS 276-1101
CR4	5 V ½ watt zener diode 1N4733 RS 276-565
R1	25 K ¼ watt potentiometer (RS 271-336)
R2	100 ohm ¼ watt (see text)
R3	510 ohm ¼ watt
R4	510 ohm ½ watt
LED	RS 276-032
printed circuit board (RS 276-024)	
LM567	phase-locked loop (RS 176-1721) (XR-567 is identical and usable)

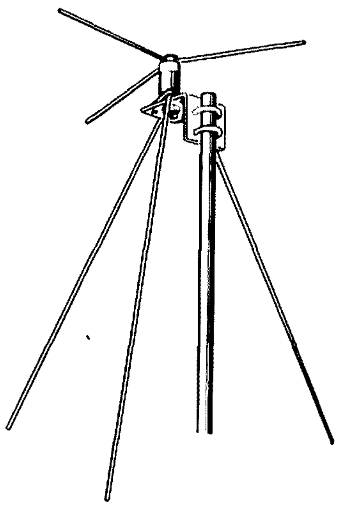
*A twisted pair made from No. 24 insulated hookup wire would work just as well and is easier to handle. **Editor**

NEW products

monitor antenna increases scanner performance

The Antenna Specialists Co. has introduced the Mon-64 Discan™, a lightweight monitor antenna that can increase base station reception by as much as 100 percent.

The Mon-64 Discan provides excellent reception on the popular low band, VHF, UHF, and T-band frequencies, from 25-512 MHz. The Mon-64 significantly improves reception of distant stations or low-powered portables. Weighing less than two pounds, the Discan antenna is easy to install, and comes complete with SO-239 connector and double U-clamp bracket; cable is not provided. The antenna mounts easily on any pipe or tubing up to 1¼ inches in diameter.

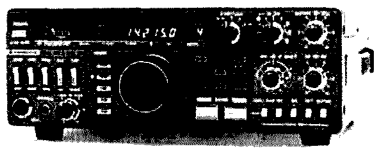


Suggested list price is $19.95. For complete specifications, contact The Antenna Specialists Co., 12435 Euclid Avenue, Cleveland, Ohio 44106. Reader Service Number 301.

TS-430S hf transceiver

The TS-430S, a recent addition to Kenwood's line of high-frequency transceivers, is an all-solid-state SSB, CW, and a-m transceiver, with fm optional. Designed to cover the 160-10 meter Amateur bands, including the new WARC bands, it also incorporates a 150 kHz-30 MHz general coverage receiver having an exceptionally wide dynamic range.

Other features include dual digital VFOs, eight memories, memory scan, programmable band scan, fluorescent tube digital display, all-mode squelch, VOX, speech processor, i-f shift, notch, and a narrow-wide filter selector



switch for use with various optional filter combinations. The TS-430S carries a factory suggested retail price of $899.95.

For additional information, contact your local Kenwood Amateur Radio dealer, or write to Trio-Kenwood Communications, 1111 West Walnut Street, Compton, California 90220. Reader Service Number 302.

C22 and B23 HT power amplifiers upgrade

Mirage Communications' pocket-size B23 (2 meters) and C22 (220 MHz) power amplifiers have been upgraded for greater versatility. A new power switch permits selection of full amplifier power or a non-energized bypass mode when only HT power is desired. The fm/SSB switch controls choice of rapid or delayed relay action. Power amplification is linear in either mode. The B23 produces 30 watts (minimum)

for 2 watts in, 15 watts for 1 watt, etc. The C22 produces 20 watts (minimum) for 2 watts in, 10 watts for 1 watt, etc. The duty cycle of both amplifiers is continuous. The B23 and C22 are made in the U.S., and carry a five-year general warranty with one year for rf power transistors.

For more information, contact Everett L. Gracey, Director of Marketing, KLM Electronics, Inc., P.O. Box 816, Morgan Hill, California 95037. Reader Service Number 303.

automatic antenna coupler

A new automatic antenna coupler, the Model H-402CU, has been introduced by Hull Electronics Company, Inc., of San Diego, California. The coupler employs a microcomputer to exactly match the ship's antenna system to any SSB radiotelephone in the frequency range of 2-22 MHz. The H-402CU initially tunes itself each time a new channel is selected and also fine tunes the antenna as necessary whenever the transmitter is on the air.

Rated at 250 watts power, the coupler combines an L network with high-efficiency toroid inductors to provide maximum transfer of the transmitter energy to the antenna. Initial tune-up time is typically ½ to 2 seconds. A unique test panel is included to allow the technician to observe operation of the coupler and to test the various digital circuits.

Installation is simple and requires no preliminary adjustment to the coupler. The H-402CU operates with antennas from 8 to 80 feet in length. Two antennas may be used with the system as an optional feature.

Model H-402CU is housed in a rugged, weather-tight enclosure 16½ inches high × 11¼ inches wide × 5½ inches deep and weighs only 9½ pounds. Operating voltage of 12 Vdc is supplied via the SSB transceiver.

For more information, contact Hull Electronics Company, Inc., 7563 Convoy Court, San Diego, California 92111. Reader Service Number 304.

noise bridge with built-in range extender

MFJ Enterprises, Inc.'s MFJ-202B noise bridge allows quick adjustment for maximum performance of any antenna — single, multiband, dipole, inverted vee, beam, vertical whip, or random systems. You can measure resonant frequency, radiation resistance and reactance. It tells you whether to lengthen or shorten your antenna for minimum SWR over any portion of the band.

The MFJ-202B will measure resistance to 250 ohms and has a wide capacitance range of ± 150 pF. It includes a built-in range extender that shunts large unknown impedances down to its measuring range. You can tune transmatches, adjust tuned circuits, measure inductance, rf impedance of amplifiers, baluns, transformers and other rf circuits. It can also be used to determine electrical length, velocity factor and impedance of coax cable. With a transmatch and dummy load, it can synthesize rf impedances.

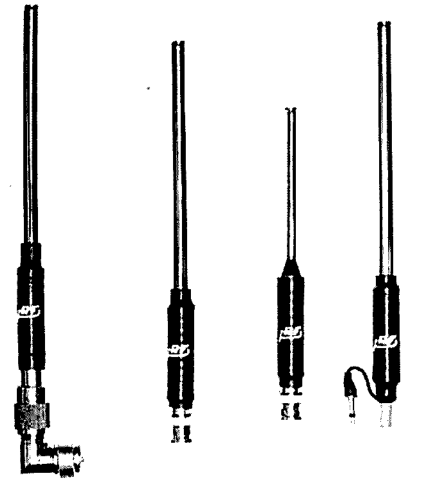
The MFJ-202B measures 4-1/2 × 2 × 4-1/2 inches and is housed in a rugged black aluminum cabinet with eggshell white front. It is available from MFJ Enterprises for $59.95, plus

$4.00 shipping and handling. For more information, contact MFJ Enterprises, Inc., P.O Box 494, Mississippi State, Mississippi 39762. Reader Service Number 305.

5/8-wave UHF antenna for handhelds

RF Products announces the addition of UHF to its existing line of 5/8-wavelength VHF telescopic gain antennas for handheld transceivers. The new models are available with a BNC type connector in 10-MHz frequency segments for the 440-512 MHz band; the most popular are now in production along with the 144-174 and 220-255 MHz versions. Typical gain is 6 dB (ref. 1/4-wave helical) or 3 dB (ref. 1/4 wave). Maximum gain and minimum VSWR is achieved by a tunable LC network.

The antennas include a base spring to prevent whip damage to the telescopic radiator. Minimum bandwidth for 1.5:1 VSWR is 10 MHz with a maximum rf power rating of 5 watts. The maximum extended length with connector is 17-3/16 inches (435 mm) and the collapsed length is 6-5/16 inches (160 mm). The operating frequency range for each model is identified by the color of the base spring cover. The model/frequency ranges available are as follows: 191-914 (440-450 MHz), 191-954

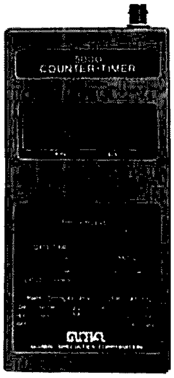


(450-460 MHz), 191-964 (460-470 MHz), 191-974 (470-480 MHz), 191-984 (480-490 MHz), 191-994 (490-500 MHz), and 191-904 (500-512 MHz). Suggested list price for all models is $19.95 with dealer and OEM discounts available.

For more information, contact RF Products, P.O. Box 33, Rockledge, Florida 32955. Reader Service Number 306.

handheld counter-timer

The 5000 Counter-Timer combines all the important features and performance capabilities of a benchtop unit with the convenience of a fully portable, battery operated instrument. It is priced at $349.95, and measures 7.6 × 3.75 × 1.7 inches, and weighs 14 ounces (without batteries).



The 5000 is designed to measure frequency, period and pulse width with extreme accuracy and exceptional reliability. It features full signal conditioning, including attenuator (X1, X10, X100); slope selection (+ or − edge for pulse-width measurement); ac or dc coupling and variable-trigger level.

A high contrast 0.43-inch LCD display offers eight-digit precision for fast and accurate readings. LCD annunciators indicate overflow, gate open, and low battery conditions. A switch allows the display storage mode to maintain the last reading in the display indefinitely.

The 5000 has automatic master reset logic, which instantly clears the display and initiates a new measurement cycle, eliminating erroneous partial measurement. A self-diagnostic function performs analysis of internal logic and provides instant assurance of accurate operation.

The 5000 has three modes of operation: frequency, period and pulse width. Signal input is via BNC connector — input impedance is 1 megohm at 25 pF for all modes. In the frequency mode, the 5000 can handle inputs from 0.1 Hz to 50 MHz. Gate times of 0.01, 0.1, 1.0, or 10 seconds can be selected. Frequency will be displayed in kilohertz on the LCD screen. The 5000 will measure any periods from 25 ns to 10 seconds and deliver a single cycle measurement or an average of 10, 100, or 1000 cycles. Time will be displayed in mS. Pulse width measurement from 25 ns to 10 seconds can be made. Either the high or low portion of the input signal can be selected.

The 5000 is powered by six AA NiCd or alkaline batteries or an optional ac adapter/charger. Optional accessories are available for the 5000.

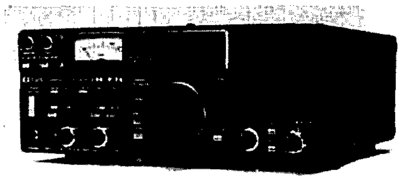
NEW products

For more information, contact Global Specialties Corporation, 70 Fulton Terrace, P.O. Box 1942, New Haven, Connecticut 06509-1942. Reader Service Number 307.

the IC-R70

ICOM's professional general coverage receiver, the IC-R70, is a generation later and features more functions than other less sophisticated general coverage receivers on the market, such as squelch on sideband, adjustable width noise blanker, adjustable speed AGC, passband tuning as standard, and adjustable notch filter as standard.

Other features are high stability, synthesized tuning and three tuning speeds, optional am/fm mode, variable CW filter widths, dial lock, two VFOs with data transfer, plus many others. The IC-R70 will operate transceive with the IC-720A, making an ideal combination for the serious DXer or CW buff.



Retail price is set at $749.00. For more information, contact ICOM, 2112 116th Ave. N.E., Bellevue, Washington 98004. Reader Service Number 308.

Moscow Muffler™ woodpecker noise blanker

The AEA model WB-1 Moscow Muffler Woodpecker Blanker offers effective blanking of the Russian Woodpecker signal with no modification to the receiver required. The WB-1 is designed to be connected in the antenna feedline between the antenna and the receiver. The WB-1 blanks the interfering pulses before they have been stretched out by receiver-tuned circuits, thereby causing the least amount of distortion possible.

Because the WB-1 is a synchronous blanker, it simply does not overload from strong adjacent channel signals. In addition to the superior blanking features, the WB-1 offers an effective low noise, broadbanded 6 dB rf preamp with +13 dBm intercept point. The preamplifier may be switched in or out whether or not the WB-1 is in the blanking mode. The WB-1 features a pulse blanking width control for reduc-

NEW products

ing the blanking width to the minimum width necessary to achieve maximum blanking. The minimum blanking width will assure the minimum signal distortion that must result from placing holes in the received signal. Blanking of both 10 Hz and 16 Hz Woodpecker modes is achieved with the WB-1.

The WB-1 Moscow Muffler Blanker is available in a transceiver version (model WB-1C) which features a carrier operated relay (COR) for automatic transfer from receive to transmit. A COR ADJ. control is provided for adjusting the relay dropout delay in switching from transmit to receive.

For more information, contact Advanced Electronic Applications, Inc., P.O. Box C2160, Bldg. O&P – 2006 196th SW, Lynnwood, Washington 98036-0918. Reader Service Number 309.

multicore emergency solder

Emergency Solder can be easily carried in a shirt pocket or stored flat, and requires only an ordinary match or candle flame to melt the solder strip. Multiple cores of rosin flux are incorporated into the flat strips, eliminating any requirement for a separate fluxing application. The flux is non-corrosive and non-conductive, and need not be removed after soldering.

To solder two wires, simply twist the wires together, wrap the solder strip lightly around them and apply a flame. Move the flame slowly back and forth until the solder flows into the splice. For larger wires, wrap two layers around the splice and use a candle to apply the flame for sustained heat. Insulating tape or sleeving should be used after soldering electrical wires.

To solder sheet metal, the solder should be placed between or on the metal parts to be connected. Hold the parts together while applying heat from a candle flame or soldering iron and then let cool. Multicore Emergency Solder is suitable for any solderable metal; it is not suitable for aluminum.

Multicore Emergency Solder costs 99 cents each. For more information, contact Multicore Solders, Cantiague Rock Road, Westbury, New York 11590. Reader Service Number 310.

learning code in stereo

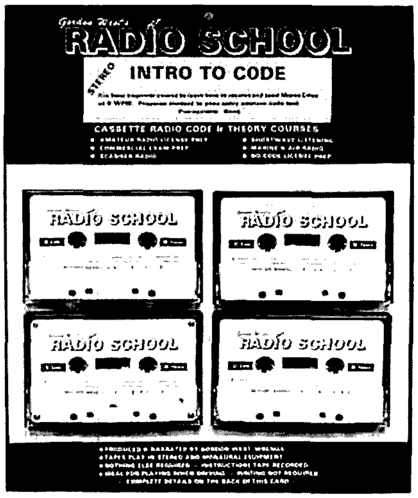
Radio School' stereo code cassettes are a new approach to learning the International Morse code. One stereo channel contains the computer-generated code which meets the latest FCC tape-speed specifications. The second channel contains the voice. Separating the

code and voice channels lets the student practice the code with or without voice help. Turning down the code channel lets students practice code letters and words in unison with the tape with their own code oscillators. Playing both channels simultaneously in a car allows students to practice the code without having to write down any letters.

The Radio School beginning/Novice code course contains four 1-1/2 hour stereo tape cassettes. All instructions are given on the cassette. By the end of the fourth cassette, students will be able to send and receive code at 6 WPM and pass a Novice class Amateur Radio code examination. They will know enough code for shortwave listening as well.

Radio School also offers code courses for upgrading to FCC General class, and to Extra class. Code test cassettes for instructors are also available. Theory cassettes are available for students wishing to pass any grade of Amateur Radio license.

The Radio School Novice stereo code course sells for $39.95. Add $3.00 for shipping. For more information, write to Radio School, 2414 College Drive, Costa Mesa, California 92626. Reader Service Number 311.



happy 30th

ANTENNA SPECIALISTS CELEBRATES ITS 30TH ANNIVERSARY

The Antenna Specialists Company got its start in 1953 by producing land mobile antennas for the Motorola and Zenith companies. Today, the Cleveland-based company is a member of the Allen Group, and it is the *world's largest and most diversified manufacturer* of communications antennas and accessories for professional land mobile services. It is also an antenna supplier for Amateur Radio, aviation, marine, personal communications, and monitor receivers.

Examples of A/S leadership are the recent introduction of the first completely concealed trunk-mounted mobile antenna for police work and the introduction of the first commercially available line of high performance base and mobile antennas for the new 800-MHz land mobile services.

In 1979, the company's Professional Division embarked on its first major departure from the antenna business with its RESCU® division, which manufactures emergency electronic devices for the public safety and industrial markets. Two years later, the firm introduced the LIFEGARD™ personal distress-alerting device for firefighters and industrial personnel working in hazardous conditions.

In 1981, A/S acquired Avanti Communications, a Chicago-based manufacturer of specialty antennas for land mobile, Amateur, autosound, and CB communications. Under the Avanti brand name, A/S has continued marketing a variety of products including CB base station antennas and an on-glass, no-ground-plane CB and land mobile antenna. The unique Avanti on-glass concept also led A/S to the development of a new mobile pager antenna system in 1982, the "Beeper Booster™."

Throughout its thirty-year history, A/S has enjoyed a reputation for product quality and reliability. Its Gold Seal Warranty on base station antennas, for example, is the only one in the industry providing reimbursement for re-installation as well as replacement of a failed antenna.

months) ($12.00 annually). Subscribe to "The Ham Boneyard", 364 Kilpatrick Ave., Port St. Lucie, FL 33452. (305) 878-7296. Master/Visa Card welcome.

EMBROIDERED EMBLEMS, ENAMELED PINS, your design, low minimum. Informational booklet. Emblems, Dept. 65, Littleton, New Hampshire 03561. (603) 444-3423.

ANTENNA — 5 band dipole/inverted V, ready to install. Send SASE for ME 5 brochure. Myad Electronics, RD1, Box 138, Linwood, NJ 08221.

SATELLITE TELEVISION INFORMATION. Build or buy your own earth station. $4.00 to Satellite Television, RD #3, Oxford, NY 13830. Parabolic antenna construction book also available. Send SASE for details.

ICOM IC-260A 2-meter synthesized all-mode transceiver. FM/CW/SSB 10 watts out. With microphone and mobile mount. Excellent condition. $325. WB8IKJ. Call 616-372-3116 nights and weekends, 616-375-6300 days; ask for Ken or leave message.

MARCONI WIRELESS TELEGRAPH CO. Stock Certificates. Authentic 1914 certificates, from the pioneering days of radio, are rare antiques and valuable investments. Suitable for framing. Only $38.95 including historical pamphlet. Satisfaction Guaranteed. Free information: Tarlen, Box 7554-M, N. Kansas City, MO 64116.

HEATH HW-101 transceiver with AC power supply. New finals, aligned, and checked. Only $350. Heath HR-1680 HF receiver. Solid-state, excellent condition. $150. WB8IKJ. Call 616-372-3116 nights and weekends, 616-375-6300 days; ask for Ken or leave message.

MOBILE IGNITION SHIELDING. Estes Engineering, 930 Marine Dr., Port Angeles, WA 98362.

WANTED, Military surplus radios. We need Collins 618T, ARC-94, ARC-102, MRC-95, HF105, VC-102, RT-804A/APN-171, RT-712/ARC-105, ARC-114, ARC-115, RT-823/ARC-131 or FM-622, RT-857/ARC-134, or Wilcox 807A, RT-859/APX-72, 313V-1 control, antenna couplers 490T, CU-1658A, CU-1669A, 490B-1, 690D-1, top dollar paid or trade for new Amateur gear, write or phone Bill Slep 704-524-7519, Slep Electronics Company, HWY 441, Otto, N. C. 28763.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. PO Box RY, Cardiff, CA 92007.

WANTED: Early production run Kenwood R-1000 (preferable early 1980 or before). Victor Barz, Rm. 3305 Cross-Baits II, Ann Arbor, MI 48109. (313) 763-2857 between 1100-1400 and 0400-0600 GMT.

MX330 Motorola factory touch tone pad with mother board and daughter board, all interconnecting wiring. Nothing to cut or glue. Complete $100.00. N6GFE, 980 Wildcat Canyon Road, Berkeley, CA 94708. (415) 843-5253.

FASTRAK® MODULAR BUILDING SYSTEM is the state-of-the-art in homebrewing. Build attractive, practical equipment for less. Send $1.00 (refundable) for info and product catalog. Proham Electronics, Inc., 34620 Lakeland Blvd., Eastlake, Ohio 44094.

MANUALS for most ham gear made 1937/1970. Send $1.00 for 18 page "Manual List", postpaid. HI-MANUALS, Box R802, Council Bluffs, Iowa 51502.

SLEP SPECIALS, HP-608FR late model signal generator 10 MHz thru 455 MHz, 19" panel with blue/gray cabinet, excellent for precision laboratory work $375.00; HP-614 signal generator 900 MHz thru 2100 MHz $345.00; SG-557/URM-52 signal generator 3.8 GHz thru 7.8 GHz $345.00; Motorola T-1034 FM signal generator, all FM bands $475.00; URM-25 signal generator 10 kHz thru 50 MHz $285.00; Tektronix IL30 spectrum analyzer plug-in 925 MHz thru 10.5 GHz $495.00; Jennings UCSL-1000 vacuum variable capacitors 5000 KV with gear drive train and mounting bracket $59.50. All lab tested excellent condition. We accept M/C, Visa or check, phone 704-524-7519, Slep Electronics Co., Hwy 441, Otto, N. C. 28763.

SATELLITE TELEVISION — Howard/Coleman boards to build your own receiver. For more information write: Robert Coleman, Rt. 3, Box 58-AHR, Travelers Rest, SC 29690.

WANTED: Schematics-Rider, Sams or other early publications. Scaramella, P.O. Box 1, Woonsocket, RI 02895-0001.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

SATELLITE TELEVISION: Discount prices on all major TVRO items. Communications Consultants. (501) 452-3149.

RUBBER STAMPS: 3 lines $3.25 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

WANTED: New or used MS and coaxial connectors, synchros, tubes, components, military surplus equipment. Bill Williams, PO #7057, Norfolk, VA 23509

VERY In-ter-est-ing! Next 5 issues $2 Ham Trader "Yellow Sheets". POB356, Wheaton, IL 60189.

CB TO 10 METER PROFESSIONALS: Your rig or buy ours — AM/FM/SSB/CW. Certified Communications, 4138 So. Ferris, Fremont, Michigan 49412; (616) 924-4561.

HAMS FOR CHRIST — Reach other Hams with a Gospel Tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

"WEST COAST 160 METER BULLETIN" devoted to, for and by top band operators. Edited and published 6 times a year by N7CKD. Subscriptions $7.00 U.S. and U.S. Possessions, Canada and Mexico $7.00 U.S. Overseas rate $8.50 U.S. dollars. To 4248 "A" St. S.E., Box 609, Auburn, WA 98002.

Coming Events

ACTIVITIES

"Places to go..."

ILLINOIS: The 1983 LAMARSFEST, Sunday, March 27, Lake County Fairgrounds, Routes 45 and 120, Grayslake. Public admitted 8 AM. Tickets $2.00 advance, $3.00 door. 9' tables $5.00 each. Free parking. Refreshments available. Talk in on 147.63-03 or 146.94 simplex. For tickets, table reservations, exhibitor information SASE to LAMARS, PO Box 751, Libertyville, IL 60048.

INDIANA: The Martinsville Hamfest, March 13, sponsored by the Morgan County Amateur Radio Club, Morgan County 4-H Building and Fairgrounds. Admission: $4.00 at door, $3.00 advance, children 11 and under free. Flea market with table $5.00; without table $3.00; premium table $20.00. Free parking. Doors open to public 8 AM. Vendor setup starts 5 AM. Talk in on 147.66/06. For tickets, table reservations and information SASE to Aileen Scales, KA9MBK, 3142 Market Place, Bloomington, Indiana 47401.

MARYLAND: The Baltimore Amateur Radio Club's 1983 Greater Baltimore Hamboree and Computerfest, March 27, Maryland State Fairgrounds Exhibition Complex, Timonium. Gates open 8 AM. Admission $3.00, children under 12 free. Large indoor dealer and flea market area. Large outdoor tailgate area. Refreshments, free parking. Guest speakers include Vic Clark, President ARRL. For information and table reservations: G B H & C, PO Box 95, Timonium, MD 21093-0095. (301)561-1282.

MINNESOTA: The 6th annual Rochester Area Hamfest, Saturday, April 9, 8:30 AM, John Adams Junior High School, 1525 NW 31 Street, Rochester. Large indoor flea market, refreshments, free parking. Talk in on 146.22/82. For further information: RARC, c/o WB0YEE, 2253 Nordic Ct. N.W., Rochester, MN 55901.

MISSOURI: The J.B.A.R.C.'s Amateur Radio Auction, March 11, Carondelet Sunday Morning Athletic Club, 1012 Loughborough, St. Louis. Open 6 PM; auction 7:30 PM. Call in on 146.94/34.

NEW HAMPSHIRE: The 3rd annual Hamfest/Flea Market sponsored by the Great Bay Radio Association, Saturday, April 9, Somersworth Armory, Somersworth, 9 AM to 3 PM. Entrance fee $1.00 per person. Refreshments available. Free parking. For advance registrations and further information: Great Bay Radio Association, PO Box 911, Dover, NH 03820.

NEW JERSEY: The Delaware Valley Radio Association's 11th annual flea market, Sunday, March 13, 8 AM to 4 PM, New Jersey National Guard 112th Field Artillery Armory, Eggerts Crossing Road, Lawrence Township. Advance registration $2.50, $3.00 door. Indoor/outdoor flea market area. Refreshments. Sellers bring own tables. Talk in on 146.52 and 146.07-.67. For information: D.V.R.A., PO Box 7024, West Trenton, NJ 08628. (SASE please.)

NEW JERSEY: The Chestnut Ridge Radio Club's Ham Radio Flea Market, Saturday, March 19, Education Building, Saddle River Reformed Church, East Saddle River Road at Weiss Road, Upper Saddle River. Tables: $10.00 for first, $5.00 each additional. Tailgating $5.00. Refresh-

ments. Free admission. Contact: Jack Meagher, W2EHD (201) 768-8360 or Roger Soderman, KW2U, (201) 666-2430.

OHIO: The TEAYS ARC will hold its sixth annual "King of the Pumpkin Hamfest," Sunday, March 20, 8 AM to 4 PM, Pickaway County Fairgrounds Coliseum. Tickets $2.00 advance, $3.00 door. 8' tables $4.00 advance, $5.00 door. Talk in .52-.52 and 147.78-18. Open for setup Saturday 4 PM. Overnight security provided. For information SASE to Dan Grant, W8UCF, 22150 Hulse Road, Circleville, Ohio 43113. (614) 474-6305.

OHIO: The Lake County Amateur Radio Association's fifth annual Hamfest and Computer Fest, Sunday, March 27, Madison High School, Madison. All indoors. 8 AM to 4 PM. Admission: $2.50 advance and $3.50 door. Table/display space $5 per 6' table; $6.50 per 8' table. Talk in on 147.81/.21. For information and reservations SASE to Lake County Hamfest Committee, 37778 Lake Shore Blvd., Eastlake, Ohio 44094. (216) 953-9784.

OHIO: The 14th annual B*A*S*H Friday night of Dayton Hamvention, April 29, Convention Center, Main and Fifth Streets. Adjacent parking. Free admission. Refreshments and entertainment. Two exciting top awards and more. For further information contact the Miami Valley FM Association, PO Box 263, Dayton, Ohio 45401.

PENNSYLVANIA: The Penn Wireless Association's Tradefest '83, Sunday, March 27, National Guard Armory, Southampton Road and Roosevelt Blvd., Philadelphia. Sellers space (6X8) $5.00. Bring tables, limited number of power connections, $3.00. General admission $3.00. Refreshments, displays and surprises. Talk-in on 146.115/715 and .52. Contact: Mark Pierson, KB3NE, 12517 Nanton Drive, Philadelphia, PA 19154.

PENNSYLVANIA: The Conemaugh Valley Amateur Radio Club's sixth annual Hamfest, Sunday, March 27, East Taylor Fire Hall, Route 271, 4 miles north of Johnstown. 8 AM to 4 PM. Plenty of food and refreshments available. Check in on the 146.34/94 repeater.

ROCHESTER HAMFEST: Atlantic Division/New York State Convention. Saturday, May 21, Monroe County Fairgrounds. Hotel headquarters, Rochester Marriott Thruway. More info? Write or call Rochester Hamfest, 300 White Spruce Blvd., Rochester, NY 14623 (716) 424-7184.

WISCONSIN: The Tri-County Amateur Radio Club's annual Hamfest, March 20, 8 AM to 3 PM, Jefferson County Fairgrounds, Jefferson. No price increase! Tickets $2.50 advance, $3.00 door. Tables $2.50 advance, $3.50 door. Free parking and plenty of food and refreshments. Talk in on 146.52, 146.22/82 and 144.89/145.49. For information, tickets and tables SASE to Horace Hilker, K9LJM, PO Box 204, 261 E. High Street, Milton, WI 53563.

OPERATING EVENTS

"Things to do..."

MARCH 12: The Green Mountain Wireless Society will operate N1VT from 1400Z to 2100Z from the Paul P. Harris Memorial Building, the site where the founder of Rotary International attended school. Frequencies: 7.235 and 21.360. Certificate available for QSO number. QSL and large SASE to Wallingford Rotary Club, PO Box 456, Wallingford, Vermont 05773, att: Ted Lidstone.

MARCH 19: B.A.R.T.G. Spring RTTY Contest. 0200 GMT Saturday, March 19 until 0200 GMT Monday, March 21. Total contest period is 48 hours but not more than 30 hours of operation is permitted. Bands: 3.5, 7.0, 14.0, 21.0 and 28 MHz. Stations: May not be contacted more than once on any one band but additional contacts may be made with the same station if a different band is used. Messages: Time GMT, RST and contact number. All logs must be received by May 31, 1983 to qualify. Summary and log sheets available from contest manager for two IRC's: Ted Double G8CDW, 89, Linden Gardens, Enfield, Middlesex, England EN1 4DX.

MARCH 19: The 13th annual Tennessee QSO party, sponsored by the Tennessee Council of Amateur Radio Clubs from 2100Z March 19 to 0500Z March 20 and 1400-2200Z March 20. Exchange signal report and county. Out-of-state stations send signal report and state, province or country. Suggested frequencies: CW - 1815 kHz and approx. 50 kHz from bottom of HF bands. Phone: 1860, 3980, 7280, 14280, 21380, 28580 kHz. Novice: 3725, 7125, 21125, 28125 kHz. Logs must be postmarked no later than May 1, 1983. Send business SASE with your log for complete results and any certificates earned to Oak Ridge Amateur Radio Club, Att: Contest Coordinator, PO Box 291, Oak Ridge, Tennessee 37830.

MARCH 26: The 4th annual Spring VHF/UHF QSO Party, sponsored by the Ramapo Mountain Amateur Radio Club, from 2100 UTC Saturday, March 26, to 0400 UTC Sunday, March 27. The grid square and range scoring system is being used. SASE to RMARC, PO Box 364, Oakland, NJ 07436 for log/entry forms and other information.

APRIL 1983 / $2.50

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hr

focus on communications technology

ham radio magazine

APRIL 1983

volume 16, number 4

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contents

REFLECTIONS

On January 20, 1983, the FCC proposed the most important Amateur rules change in many years — the proposal to delegate the responsibility for Amateur license examinations to the Amateur community. Unfortunately, the emotion-laden issue of a no-code license has all but obscured this other crucial Notice of Proposed Rule Making.

In consideration of the ARRL's Petition for Rule Making on exam administration, the FCC has proposed a three-man team headed by an Extra Class licensee to administer individual exams. Examiners would have to be certified by one of several recognized supervisory organizations called Volunteer Examiner Coordinators. They would have to be over 18 years old and could not work for a manufacturer or distributor of Amateur equipment or a publisher of training materials. Questions would be chosen from a list made up by the Commission from submissions by individual Amateurs and groups of Amateurs.

There are a lot of hard questions that must be asked about this proposed system. Three-man examining teams (for all licenses except Novice) are fine for major urban areas like Los Angeles, Chicago, or Washington, D.C., but what happens to the would-be upgrader in remote parts of the country? Should there be a mechanism provided to deal with such cases, for example an examining team led by an Amateur but including non-Amateur examiners, such as elected public officials, when three licensed Amateurs aren't available? Should a formal procedure for giving exams at hamfests or in classrooms be included in the new rules?

What should the qualifications for a Volunteer-Examiner-Coordinator (VEC) be? There has been definite interest in this program shown by some non-Amateur groups. How should the long-term integrity of the VECs be ensured?

It seems that the FCC would prefer to have more than one VEC overseeing the exam-administration effort. How could anyone be sure that the different organizations all hew to the same standards? How would the overseeing groups finance their administration costs? The ARRL is already well aware of what this program is going to cost it, and it questions whether it's fair to the League's members to have them pick up the bill. Should there be a fee charged for giving Amateur exams? Who should set the fee, and to whom — the examiners, their overseeing group, or both — should it go?

Should the FCC include the Novice exam in this new overall program, instead of establishing the less demanding Novice exam program they proposed in an NPRM late last year? The ARRL wants Novices included, yet the Commission has indicated its approach would be simpler, faster, and cheaper. The Commission received very few comments on its Novice exam NPRM; does that mean Amateurs want the Novice exam a part of the larger program, or was the FCC's proposed Novice exam program simply overlooked in the concern generated by the no-code license proposal?

There are other considerations as well. It takes time to establish workable procedures (look how long the FCC had). Might not inordinate delays occur at every step of the process, resulting in longer delays in getting licensed? Right now it's a hot topic, but what about one or two years downstream? Might not interest wane among the exam administrators — with newcomers to the hobby being the losers? Most of all, we should be concerned about the possibility that the ham ticket might be devaluated by an unequitable, non-uniform examination procedure. How simple it seems now, to go down to the nearest FCC office and take the exam. Might not a small licensing fee underwrite the cost of FCC-administered exams?

Write the FCC with your opinions. Comments on the exam administration proposal, FCC PR Docket 83-27, are due at the Commission by April 8th. Address them to the Secretary, Federal Communications Commission, Washington, D.C. 20554. You'll need to send an original neatly typed with wide margins, plus five copies (eleven is better, since each Commissioner will receive one). Your name and the Docket number should appear on each page.

What we, as individual Amateurs and through our clubs and organizations, tell the Commissioners may do more to influence the future of Amateur Radio in the United States than anything else we will ever do!

ham radio

comments

filters for Amateur use

Dear HR:

In his letter to the *ham radio* editor (February, 1983, page 8), Ed Marriner, W6XM, mentioned a problem the Radio Amateur too frequently ignores — the need to comply with the FCC requirement that transmitter harmonics be down by 40 dB or more from the fundamental. Ed further explained that to accomplish this on all bands a lowpass filter for each band is necessary. The customarily used 30-MHz lowpass filter, widely advertised by J.W. Miller, Drake, and B&W, is effective, he said only for the Amateur 10-meter band.

A "best solution" offered by Ed was for the Amateur to install lowpass filters designed to cut off just above the upper end of the band being used; however, the recommended designs were from the June, 1957, issue of *GE Ham News* — designs that are more than twenty-five years old!

During the past twenty-five years, the Radio Amateur has witnessed many changes, the most obvious being the transition from vacuum tube to solid state, and more recently the introduction of the personal computer to ham operation. Less obvious was the transition from filter design using the image-parameter-design procedure invented by Otto Zobel to the modern filter (network-synthesis) design procedure. The modern design filter has a simpler configuration and a more precise performance than the older image-parameter type. Modern lowpass designs (Chebyshev and elliptic) have been developed in which standard-value capacitors are used, thus making them simple for the Amateur to build. These designs have been widely published in the Amateur Radio handbooks, in trade handbooks, and in the Amateur and trade periodicals. I think Ed will agree that these designs are a better solution to the Amateur lowpass filtering requirements than are the old designs.

Ed also mentioned the problem of obtaining suitable high-voltage, low-loss capacitors for use in constructing lowpass filters for Amateur high-power applications. I, too, have experienced this problem, and I have continually been searching for a better high-voltage capacitor than the Centralab ceramic TVL type that Ed mentioned. I think I have finally found a suitable alternative to the TVL capacitor, but the manufacturer of the high-voltage capacitor, KD Components Inc. (3016 S. Orange Ave., Santa Ana, California 92707), sells only in quantities greater than ten and has a minimum billing of $50. Also, the maximum capacitance available in the 2-3 kV range is 100 pF, so several capacitors will have to be paralleled to get the larger capacities required by the filters for the lower Amateur bands. The approximate cost of the 2-kV, 100-pF, 10 percent capacitor in quantities of 10-99 is $4. In quantities above 500, the price drops to $1.44! Consequently, this capacitor type, although excellent for the application, appears to be financially practical only for a high-volume manufacturer of lowpass filters.

A filter designed from the data in reference 4 (*QST*, December, 1979) was constructed and operated at a 1-kW power level without a failure, but this is feasible only when the VSWR can be carefully controlled, otherwise the voltage rating of the capacitors may be exceeded and the filter damaged if the VSWR becomes excessive. For power levels below 500 watts, the polystyrene and mica capacitors seem suitable. So, contrary to Ed's concluding statement, there does seem to be hope, and I suggest that those not having a filter for each band should review the references included with this letter, and then construct any filters that may be required.

references

1. *Radio Handbook*, 22nd edition, edited by W.I. Orr, W6SAI, pages 3.35-3.37, copyright 1981 by H.W. Sams & Co., Inc., Indianapolis, Indiana.
2. *The Radio Amateur's Handbook* (1983), 60th edition, edited by George Woodward, W1RN, page 6-41, fig. 65, copyright 1982 by ARRL, Newington, Connecticut.
3. *The Electronic Databook*, edited by R. Graf, to be published in 1983 by TAB Books, Blue Ridge Summit, Pennsylvania.
4. Ed Wetherhold, W3NQN, "Low-pass filters for amateur radio transmitters," *QST*, December, 1979.
5. Ed Wetherhold, W3NQN, "Elliptic lowpass filters for transistor amplifiers," *ham radio*, January, 1981.
6. Ed Wetherhold, W3NQN, "Filter Design," Comments, *ham radio*, October, 1981, page 6.
7. Ed Wetherhold, W3NQN, "7-element 50-ohm Chebyshev filters using standard-value capacitors," *rf design*, Vol. 3, No. 2, February, 1980, pages 26-38.
8. Ed Wetherhold, W3NQN, "Low-pass filters (with inductive input and output)," *rf design*, Vol. 4, Nos. 4 and 5, July/August and September/October, 1981.
9. Ed Wetherhold, W3NQN, "Design 7-element lowpass filters using standard-value capacitors," *EDN*, Vol. 26, No. 1, January 7, 1981, pages 187-190.
10. Ed Wetherhold, W3NQN, "Low-pass Chebyshev filters use standard-value capacitors," *Electronics*, Engineers' notebook, June 19, 1980.

Ed Wetherhold, W3NQN
Annapolis, Maryland

presstop de W9JUV

LEGAL PROBLEMS WITH BOTH ANTENNAS AND RFI are continuing for a number of Amateurs in various communities, and should be of concern to Amateurs throughout the country. Cerritos, California, long a leader in antenna height restrictions, has resisted a suit brought by several Amateurs attempting to overturn a moratorium on new antennas. That city is in the process of drafting a new, more restrictive ordinance, but the judge ruled the suit was "premature," since the terms of the new ordinance haven't yet been decided. An appeal is being considered, and the ARRL Executive Committee has voted in favor of entertaining a request for the League to match local financial support. Other tower cases in Oklahoma City (N5SW), Farmington, Michigan (WD8BCM), and Burbank, Illinois, are still pending, but NØBCX's challenge of a Brighton, Minnesota, ordinance has been upheld and his 65-footer is still up.

WB2BZK's Appeal To The New Jersey Supreme Court of Winslow township's prohibition of RFI has been turned down. Efforts are continuing to get the township to review and perhaps rescind its ordinance, in view of the federal assumption of such regulation contained in Public Law 97-259 (the "Goldwater Bill").

TWO VITALLY IMPORTANT AMATEUR RULES CHANGES PROPOSED BY THE FCC are up for comments during April. The proposal to establish a volunteer examining procedure for all Amateur licenses, PR Docket 83-27, has a comment closing date of April 8, which leaves little time to consider its implications (see this month's Reflections, page 6). The no-code license proposal, PR Docket 83-28, is open for comments until April 29.

The ARRL's Adamant Position Against No-Code May Be Softening, according to some indications. It appears that the League membership may not be as solidly against a no-code license as was previously reported. With the strong pro no-code position apparent among the Commissioners, the ARRL now feels it may be prudent to support a form of no-code that the Amateur community can live with, rather than oppose it outright and have no say in its final form. The League Executive Committee has agreed informally to prepare a position paper outlining various no-code alternatives for the directors before their April board meeting, at which time the final League position will be determined and comments prepared.

Amateur Exam Administration At This Year's Dayton Hamvention had been sought by the Hamvention Committee, and initial FCC reaction had been positive. However, it now appears that, though exams will be given at the Hamvention, they will be administered under the supervision of the FCC. Barring unforeseen problems, the earliest a complete volunteer program could be put together and set in motion would be late next fall, leaving too many variables to be settled in time for even a dry run at this year's Hamvention.

RICH ROSEN, K2RR, HAS BEEN APPOINTED EDITOR-IN-CHIEF OF HAM RADIO effective February 5. Rich joined Ham Radio last fall, as Senior Technical Editor.

ARRL's New Technical Department Manager Is Paul Rinaldo, W4RI, who's replacing Doug DeMaw, W1FB, upon Doug's retirement in May. Paul currently edits QEX, the ARRL experimenters' newsletter, and is the president of AMRAD.

TEN SCHOLARSHIPS FOR GENERAL CLASS (OR HIGHER) AMATEURS planning to attend (or already attending) college or technical school are available through the Foundation for Amateur Radio. Full details and an application form can be obtained from Hugh Turnbull, W3ABC, 6903 Rhode Island Ave., College Park, MD 20740. May 31 is closing date for requests.

THE PHONE BAND EXPANSION IS STILL IN PROCESS within the Commission, with expectations that it will be finished and released by late spring. Just which bands will (and which won't) be changed isn't yet clear, though it seems very likely that 20 and 10 will both see some expansion of their phone subbands.

Deregulation Of The CB Service Is Also In The Mill, with the new rules (or non-rules) to be announced at about the same time as Amateur phone band expansion.

Extension Of Amateur License Terms To 10 Years from the present five is likely to surface soon. Though the FCC now has the authority to make the change, it will probably be introduced through a Notice of Proposed Rule Making to assess Amateur reaction.

PROFANE AND INDECENT LANGUAGE IS NO LONGER GROUNDS FOR REVOKING an Amateur's license, according to the FCC's Review Board. The license of N6BHU had been revoked last fall by an FCC Administrative Law Judge for such violations, but on January 26 that decision was overturned and his license reinstated. With a suburban Washington, D.C., broadcast station now airing uncensored "party" records, it appears the Commission concern with the content of transmissions may be a thing of the past. However, the ARRL is seeking a review of the subject with the FCC staff in hopes of restoring some standards for Amateurs.

THE SOLAR FLUX SLUMPED TO ITS LOWEST LEVEL SINCE JANUARY, 1978, at mid February, to give a hint of things to come as this sunspot cycle deepens. Solar activity remained low through CW DX Contest weekend, with 10 meters of little value and 15 spotty.

Deteriorating HF Band Conditions Highlight The Value of Beacons, particularly the new 14.100 MHz worldwide system sponsored by the Northern California DX Foundation. In addition, the many beacons in the 28.2-28.3 MHz portion of 10 meters and those operating between 50.0 and 50.1 MHz on 6 will continue to signal openings to users of those bands. See Technical Forum, page 46, for information on a beacon on 28.208 MHz.

inexpensive video monitor

Bypassing rf and i-f sections to resurrect old TVs for modern use

The current interest in home computers, slow-scan and fast-scan TV, RTTY, and automatic CW keyboards — not to mention home video movies and games — creates a need for an inexpensive display device for the ham shack. Many commercial video products are designed to work with a standard TV, typically using channels 2, 3, or 4 with a video modulator. There are some drawbacks to this procedure, though; for one thing, the family TV is not likely to be located in the ham shack. And, more importantly, the performance of a TV set is less than optimum if high resolution is needed.

I first considered the problems of TV sets when I acquired a TRS-80 Model I microcomputer a few years ago. I figured I could save some money by converting an old black and white TV set for use as a monitor. Typical computer-grade monitors sell for $100 or more, but a flea-market TV can be found for next to nothing. And TV sets have a 15 to 20 inch screen, unlike the typical 12-inch monitor. Sounds like a bargain, but there's a hitch.

The problem is bandwidth, (or resolution, depending on your point of view). Commercial CRTs use an 80-character display, and many home computers settle for 48, 32, or even 24 characters per line. The resolution of TV is typically much less. The TRS-80 uses a sixty-four character display, which is why Radio Shack sells a dedicated monitor. Those sixty-four characters occupy roughly 80 percent of the horizontal scan line. Each character is five dots (pixels) wide, and there is a one-pixel space between letters. So, we have *6 × 64 = 384 pixels per line*. In a conventional (U.S.) TV scanning system, one line is scanned in 63.5 microseconds. Only 80 percent of this time is available for the letters, so the pixels are scanned at a rate of *(384 pixels/line)/(0.8 × 63.5 microseconds) = 7.6 million pixels/second*, or 130 nanoseconds/pixel! The situation is even worse for an eighty-character line. (The longer lines are desirable for RTTY — where seventy-two character machines are common — and word-processing.) Furthermore, in order that the pixels reach full brightness when on, and return to the black video baseline when off, the rise and fall times must be much less than the 130-nanosecond duration of a pixel. Otherwise they will run together in the bars on the letters T, E, B, and so on, and fade out in the vertical part of letters I, L, etc., as noted by W9CGI.[1] We require a bandwidth at least twice the pixel rate, or 15 to 20 MHz!

Broadcast TV uses a 6-MHz channel width. The i-f strip is designed to have sharp cutoff, to minimize adjacent channel interference. The video carrier is already 1.25 MHz above the lower band edge in the

By Carl Gregory, K8CG, 203 Trappers Place, Charleston, West Virginia 25314

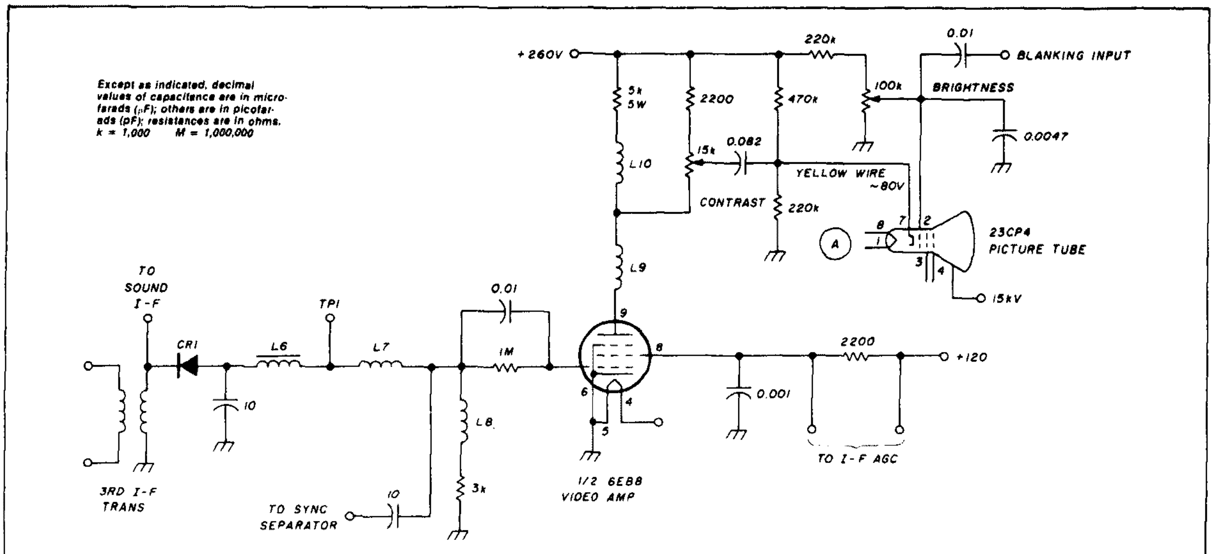

fig. 1. Partial schematic of black and white TV set showing video amplifier and picture tube connections. CR1 is a 1N60, 1N395, etc. L7-L10 are video peaking coils. Points TP1 and A are referred to elsewhere.

vestigial sideband system. The maximum available video bandwidth is a bit more than 4 MHz, if we use the video-modulated rf carrier approach. It should have been no surprise to me when my TRS-80 display was illegible unless I used the expanded (thirty-two character per line) display. The problem was compounded by snow (low signal to noise ratio) from the aging rf section.

How to remedy this? Several approaches are possible:

1. Slow down the scan rate. This method has two drawbacks in that it causes annoying flicker in the display, and requires major modifications to the TV and the video display-generating circuitry.

2. Pre-process the video signal, emphasizing high frequencies. This approach was used by W9CGI. But you can compensate for only so much high frequency roll-off. The match between compensation and i-f roll-off must be exact.

3. Bypass the problem by skipping the rf and i-f sections of the TV set. This is my approach; it is the simplest, the most effective, and potentially the cheapest, since you can use a TV with a defunct tuner.

Let's consider the modification of a typical tube-type TV set for use as a wideband video display. **Fig 1** shows the pertinent parts of the video amplifier stage. (The circuit is from my Setchell-Carlson set[3]). Note that the cathode of the picture tube is driven. **Fig. 2** shows the typical video signal level available from the computer, demodulator, or other source. Our problem is to match the two devices.

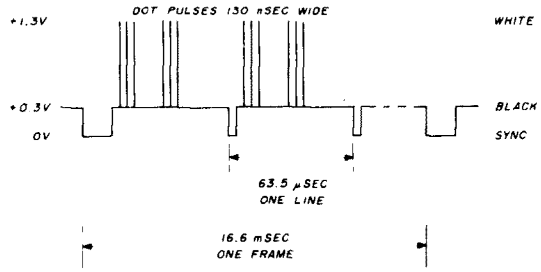

fig. 2. Typical video text display format. The actual picture is about 1.0 volt peak-to-peak with sync pulses 0.2 to 0.3 volt below black. The whole signal is usually positive, as shown. (Levels measured across 75 ohms.)

simplest approach

The first method uses the existing video amplifier and bypasses the rf and i-f stages. In my set, the detector output (TP-1) was brought out to a test point on top of the chassis. I determined that, since the typical detector level is a couple of volts peak-to-peak, the new video source could just be hooked in parallel here.

Fig. 3 shows the circuit. The signal comes from the computer, VCR, or other source, via the coax. Some 50 or 75-ohm cable is all right as long as it's not more than a few feet long. I used RG-174, which is a nice size — not too stiff. A blocking capacitor is

needed since the 2200-ohm grid leak would otherwise be shorted by the 75-ohm video source. This capacitor must pass frequencies as low as the vertical sync pulses at 60 Hz. For 50- or 75-ohm systems this means $Xc = 50\ ohms$ (maximum), so $C = 1/(2 \times pi \times 60 \times 50) = 53\ microfarads$ (minimum). Note the polarity of the capacitor: the grid is negative. *Be cautioned: This circuit will not work on a transformerless TV with a hot chassis unless an ac isolation transformer is installed,* because there is no place to safely install the shield side of the video cable!

I installed this circuit (cost was less than $3.00 for cable, connectors, and capacitor) in my TV set, and was using my new computer in a day or so. I used it that way for about a year before I had time to try to improve the performance. There was still considerable blurring of pixels, the *brightness and* contrast needed continuous adjustment as the set warmed up, and the contrast was a bit low.

second approach

A note in *Byte* magazine suggested the solution.[2] A new video amplifier improves the high-frequency response, and provides the needed gain to get the desired contrast. Unfortunately, the circuit in the magazine had some drawbacks, and I finally came up with my own. The circuit in **fig. 4** has the following properties: no separate power supply needed; adequate gain; sufficient bandwidth to give well-formed characters at 80 per line; linear enough to use for a video-tape or ATV monitor; uses the existing brightness control circuit; no exotic devices required. (Any modern PNP and NPN transistor should work fine.)

circuit description

The video cable is terminated in a resistive pad,

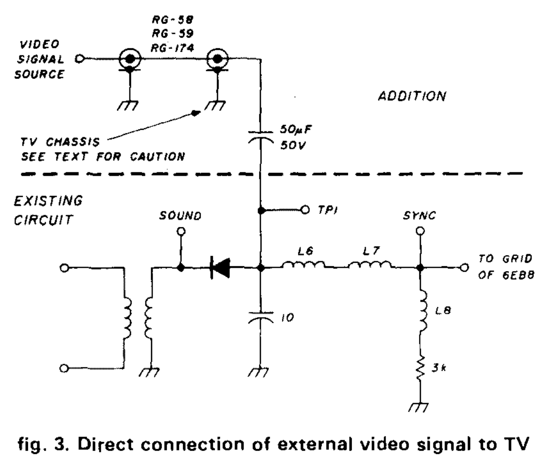

fig. 3. Direct connection of external video signal to TV set in fig. 1.

R1-R3, which also serves as part of the bias circuit for Q1. C1 passes the signal to the video amplifier (for the benefit of the sync circuitry only). Q1 operates as a linear common-emitter amplifier. The gain is about 5 at low frequencies. The emitter bypass (C2) boosts the gain above 5 MHz and compensates for the transistor's reduced beta. It is important to operate all the transistors in the linear region. Saturation of any transistor (clipping) will load the base with charge, requiring time to discharge, resulting in slow switching or reduced bandwidth.

Q2 acts as an inverter and level shifter. The gain is about 0.7. The level shifting is necessary since dc coupling is used to get wide bandwidth, and the gain need not be high since we have one more stage to go. Q3 is another common-emitter amplifier, with a gain of 3 at the collector. The signal has now been inverted three times, so it is inverted overall. The white peaks of the video input are negative peaks at the output (cathode) which drive off many more electrons and make a white spot on the screen.

The cathode of the picture tube is a low current point, so we can use a relatively small capacitor to couple the signal to it (C3). We also need dc here, but the value (+80 volts) is a bit high for the transistors, so I derived it from the existing bias circuit via a filter (R11-12 and C4). The filter removes the video information from the old amplifier, but passes the dc cathode current.

The overall gain is about −15, so a 1-volt peak-to-peak input yields about 15-volts peak-to-peak output. This means the power supply must provide considerably more than 20 volts. By trial and error, dropping resistors (R13-14) were found which gave an acceptable picture without overheating the transistors. (It is difficult to calculate the value, since the current drain of the amplifier varies with Vcc and the nature of the video signal.) C5 filters the resulting 40 volts or so and keeps it relatively constant during each frame. This unregulated supply is the least satisfactory aspect of the circuit, but the heat from the dropping resistors is hardly noticed in a tube-type set.

construction

This amplifier can be built in a breadboard format, since parts placement is not critical. I used a simple printed circuit board (cover the foil with masking tape and remove some with a knife, leaving large islands where the components are to be attached). The board was installed in the back of the set below the base of the picture tube. I just cut the yellow cathode lead in the middle (point A in **fig. 1**) and attached the two ends to the board. This is probably the only semi-critical item — it wouldn't do to go to all that trouble to get a sharp video signal and then

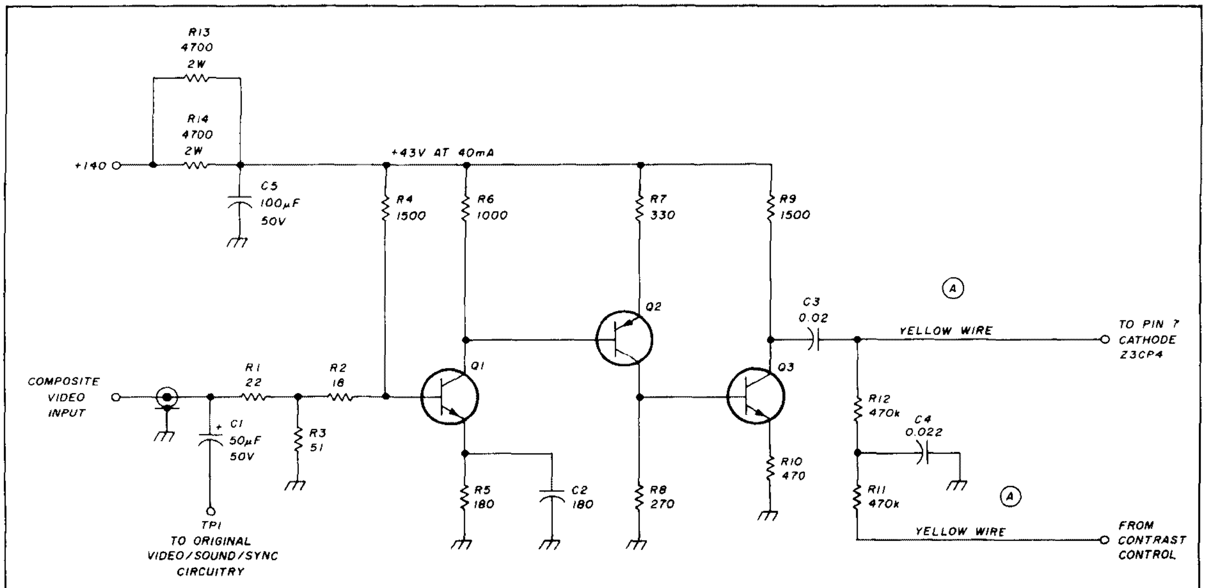

fig. 4. New broadband video amplifier for use with tube-type TV sets. For external connections refer to fig. 1. The cathode lead is cut at point A, and the amplifier is inserted as indicated. Q1, Q3 are 2N2222A and Q2 is a 2N3905.

run it through a long inductive lead. But 6 to 8 inches doesn't seem to hurt. *Caution: This circuit will not work on a transformerless TV with a hot chassis unless an ac isolation transformer is installed,* because there is no place to safely install the shield side of the video cable!

adjustment

There are no adjustments in the new video amplifier. However, we can improve performance a bit by adjusting the TV set. So far, only the video characteristics associated with the horizontal sweep have been discussed. Another problem with TV is that the scan is not necessarily linear. That is, the picture may be bunched up at one side or the top or bottom of the screen, and spread out elsewhere. This would be intolerable in a color set, but is common in old black and white sets. There should be a pair of pots for adjusting top and bottom vertical linearity, and there are two tabs mounted on the yoke which can be rotated independently to position the picture properly. It should be possible to find a combination in which the part of the picture you want to use the most is conveniently displayed.

Finally, you may find that the focus is not uniform over the screen. The focus control and those positioning tabs may interact, and it should be possible to get a reasonably crisp picture over the most important parts of the screen. (For use as a computer or RTTY display, the right side is less important than the left, for instance.)

I have been using the circuit in **fig. 4** with the TRS-80, and more recently with a homebrew S-100 system using an eighty-column display, for about a year now. It sure is nice to not have a lot of short, left-over lines on RTTY — you can see most of the last transmission all at once with the eighty-column display. The large screen is easy on the eyes, although the contrast is a bit low if the room is sunlit. And word-processing is handy, too; this article was written on the big screen. A test with a Panasonic portable VCR and camera showed an excellent picture of the shack. (Everything perfect except the color!)

An old TV set can be given a new lease on life as a modern video monitor for the shack at relatively low cost. The effort required will depend on the bandwidth needed for your application. I hope the principles presented here will save you a good deal of time and frustration as you attack that tube. This project is worthwhile for any experimenter on the more modern modes, such as RTTY, slow or fast-scan TV, or computer communications.

references

1. D.J. Brown, W9CGI, "CRT Character Enchancer," *ham radio,* August, 1982, page 66.
2. Timothy Loof, "Use Your Television Set as a Video Monitor," *Byte,* February, 1979, page 46.
3. Schematics for older TV sets can be obtained from your local TV-repair parts supplier. Ask for Howard W. Sams "Photofact" sheets by model number.

ham radio

Morse time synthesis

This software routine lets your micro speak the time in Morse code

Talking microcomputers are becoming common as more companies develop hardware modules for voice synthesis. Most are reasonably priced, starting from $100. If you want vocal feedback from your computer and need only a vocabulary limited to the decimal integers, you might consider a software alternative: voice synthesized telegraphy.

The program described here synthesizes a 24-hour clock which provides an audible read-out in Morse code characters. The clock produces Morse characters for the time in hours and minutes on demand, and automatically on the hour. It is especially useful to the blind or seeing-impaired. Even to those unfamiliar with Morse code, the numerals are easy to learn.

This program was conceived on a single-board computer based on the 1802 microprocessor, running at a frequency of 1.7897725 MHz. The program depends upon only a few hardware features: 256 bytes of RAM, a speaker amplifier on the Q line, and a push-button switch on the EF4 external flag line. All the routines in the program are straightforward, and they can be easily translated into machine code for other microprocessors.

the program

The main program begins by initializing registers to point to subroutines, locations for binary time, BCD time, and a table of Morse code patterns. The program then enters the time loop (**fig. 1**). This loop iterates until sixty seconds elapse, at which time binary minutes are incremented. During each iteration, the program checks to see whether the time has been requested by testing external flag EF4, and it checks to see whether sixty minutes have elapsed. In either case program control passes to register R3 for the out subroutine. The time loop also checks for twenty-four hours, at which point the clock is reset to 00:00.

The out routine (**fig. 2**) is the main subroutine, and produces the Morse code characters. It first clears the old BCD time. Then, after getting binary hours, the program jumps via R4 to the BCD subroutine. On return, the program gets binary minutes and again

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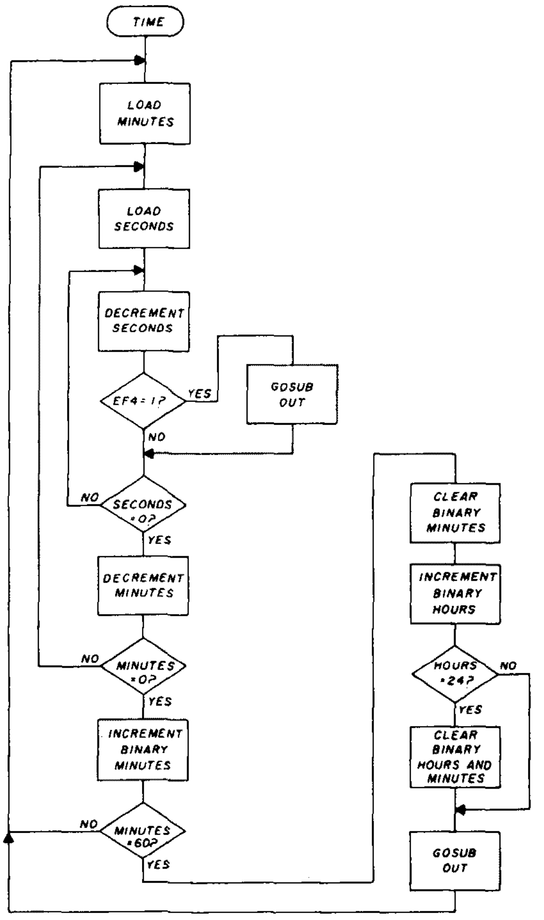


fig. 1. Flowchart of the time loop.

jumps to the BCD subroutine (**fig. 3**). Now that the program has the time in BCD, control passes to R5 for the Code routine, which converts the BCD digits into Morse characters. On return, program control reverts to R0 to resume the time loop.

BCD is a fairly conventional binary-to-BCD subroutine which converts by successively subtracting ten from the binary value. Every time it subtracts ten, it increments the BCD ten's value and retains the difference as the new binary value. If the subtraction yields a negative difference, the previous binary value is stored as the BCD one's value.

For each BCD digit the code subroutine (**fig. 4**) finds the bit pattern which corresponds to the appropriate Morse code digit. It then takes this bit pattern and ring-shifts it right, into DF (the 1802's carry flag) five times. After each shift, the code subroutine tests DF. If DF is 0, a dot is fetched; if DF is 1, it returns a dash. For instance, for the numeral 2 (·· — — —) the bit pattern fetched from the table will be XXX11100. (The higher-order three bits are not used.) The code routine will also generate a space between Morse digits.

The routine which produces the tones is called

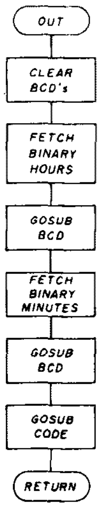


fig. 2. Flowchart of the out routine.

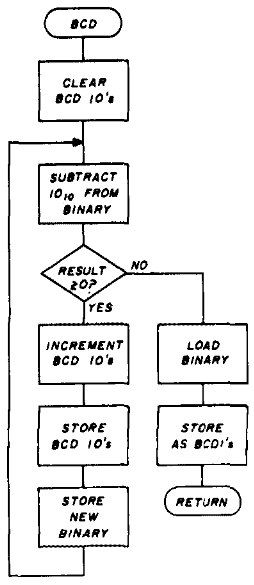


fig. 3. Flowchart of the BCD subroutine.

Program listing.

Register	Use
0	Main program counter
1	Not used
2	Not used
3	Pointer to OUT (0055)
4	Pointer to BCD (0076)
5	Pointer to CODE (0088)
6	Pointer to bit pattern for Morse digits.
7	Pointer to DOT/DASH (00B6)
8	Scratch pad
9	Pointer to Binary Hours (00D0)
A	Pointer to Binary Minutes (00D1)
B	Pointer to BCD Time (starting with 10's hours at 00D2
C	Counter for Minutes
D	Counter for Seconds
E	Morse character counter (low order)/Temporary code storage (high order)
F	Pointer to top of table of bit patterns for Morse digits (starting at 00D6)

Address	Code	Label	Mnemonic	Operand	Comment
0000	F800	INIT	LDI	00	Set high order of registers
0002	B3		PHI	R3	to 00.
0003	B4		PHI	R4	–
0004	B5		PHI	R5	–
0005	B6		PHI	R6	–
0006	B7		PHI	R7	–
0007	B9		PHI	R9	–
0008	BA		PHI	RA	–
0009	BB		PHI	RB	–
000A	BF		PHI	RF	–
000B	F855		LDI	55	R3 points to OUT (0055).
000D	A3		PLO	R3	–
000E	F876		LDI	76	R4 points to BCD (0076).
0010	A4		PLO	R4	–
0011	F888		LDI	88	R5 points to CODE (0088)
0013	A5		PLO	R5	–
0014	F8B6		LDI	B6	R7 points to DOT/DASH (00B6).
0016	A7		PLO	R7	–
0017	F8D0		LDI	D0	R9 points to Binary Hours (00D0).
0019	A9		PLO	R9	–
001A	F8D1		LDI	D1	RA points to Binary Minutes (00D1).
001C	AA		PLO	RA	–
001D	F8D2		LDI	D2	RB points to BCD Time (00D2).
001F	AB		PLO	RB	–
0020	F8D6		LDI	D6	RF points to Table of Morse
0022	AF		PLO	RF	digits.
0023	F83C	TIME	LDI	3C	Load minute count . . .
0025	AC		PLO	RC	in RC.0.
0026	F86D	SEC	LDI	6D	Load second count . . .
0028	BD		PHI	RD	in RD.1.
0029	2D	DECSEC	DEC	RD	Decrement second count.
002A	3F32		BN4	GETSEC	Anyone want the time?
002C	372C		B4		
002E	F855		LDI	55	If yes, set OUT sub pointer
0030	A3		PLO	R3	R3 to 0055 and . . .
0031	D3		SEP	R3	gosub OUT.
0032	9D	GETSEC	GHI	RD	Get second count.
0033	3A29		BNZ	DECSEC	If ≠ 0, go to DECSEC again.
0035	2C	DECMIN	DEC	RC	Decrement minute count.
0036	8C		GLO	RC	Get minute count.
0037	3A26		BNZ	SEC	If ≠ 0, go to SEC again.
0039	0A		LDN	RA	Get binary minutes.
003A	FC01		ADI	01	Increment binary minutes.
003C	5A		STR	RA	Store new binary minutes.
003D	FF3C		SMI	3C	Have 60 minutes elapsed yet?

Address	Code	Label	Mnemonic	Operand	Comment
003F	3A23		BNZ	TIME	If not, get another minute.
0041	F800		LDI	00	If yes, clear binary minutes
0043	5A		STR	RA	and . . .
0044	09		LDN	R9	increment binary hours.
0045	FC01		ADI	01	–
0047	59		STR	R9	–
0048	FF18		SMI	18	Have 24 hours elapsed yet?
004A	3A4F		BNZ	GONG	If not, output hourly gong.
004C	C4		NOP		–
004D	59		STR	R9	If yes, reset time to 00:00
004E	5A		STR	RA	–
004F	F855	GONG	LDI	55	Set OUT sub pointer R3
0051	A3		PLO	R3	to 0055 and . . .
0052	D3		SEP	R3	gosub OUT.
0053	3023		BR	TIME	Back to TIME.
0055	F800	OUT	LDI	00	Clear BCD 10's Hours.
0057	5B		STR	RB	–
0058	1B		INC	RB	Clear BCD 1's Hours.
0059	5B		STR	RB	–
005A	1B		INC	RB	Clear BCD 10's Minutes.
005B	5B		STR	RB	–
005C	1B		INC	RB	Clear BCD 1's Minutes.
005D	5B		STR	RB	–
005E	F8D2		LDI	D2	Restore BCD Time Pointer RB
0060	AB		PLO	RB	to 00D2.
0061	F876		LDI	76	Restore BCD sub pointer R4
0063	A4		PLO	R4	to 0076.
0064	09		LDN	R9	Get Binary Hours and . . .
0065	D4		SEP	R4	gosub BCD.
0066	F876		LDI	76	Restore BCD sub pointer R4
0068	A4		PLO	R4	to 0076.
0069	OA		LDN	RA	Get Binary Minutes and . . .
006A	D4		SEP	R4	gosub BCD.
006B	F8D2		LDI	D2	Restore BCD Time pointer RB
006D	AB		PLO	RB	to 00D2.
006E	F888		LDI	88	Restore CODE sub pointer R5
0070	A5		PLO	R5	to 0088.
0071	D5		SEP	R5	Gosub CODE.
0072	F8D2		LDI	D2	Restore BCD Time pointer RB
0074	AB		PLO	RB	to 00D2.
0075	DO		SEP	RO	Return to TIME.
0076	A8	BCD	PLO	R8	Put Binary in scratch pad R8.0
0077	FF0A	SUB10	SMI	OA	Subtract 10_{10} from binary.
0079	3B83		BNF	BCD1'S	Use Binary as BCD 1's if result
007B	A8		PLO	R8	≤ 0. Otherwise store result . . .
007C	OB		LDN	RB	and increment BCD 10's.
007D	FC01		ADI	01	–
007F	5B		STR	RB	–
0080	88		GLO	R8	Get new Binary and try to
0081	3077		BR	SUB10	subtract 10 again.
0083	1B	BCD1'S	INC	RB	Since Binary is less than
0084	88		GLO	R8	10 store it as BCD 1's
0085	5B		STR	RB	–
0086	1B		INC	RB	–
0087	D3		SEP	R3	Return to OUT.
0088	F830	CODE	LDI	30	Delay between Morse digits.
008A	B8		PHI	R8	–
008B	28	DECDEL	DEC	R8	Decrement delay value.
008C	98		GHI	R8	–
008D	3A8B		BNZ	DECDEL	Time up yet?
008F	8B		GLO	RB	Check to see if the last

Address	Code	Label	Mnemonic	Operand	Comment
0090	FFD6		SMI	D6	Morse digit has been output.
0092	C6		LSNZ		If it has, return to OUT.
0093	D3		SEP	R3	–
0094	C4		NOP		–
0095	OB		LDN	RB	Get BCD value.
0096	IB		INC	RB	–
0097	FCD6		ADI	D6	Add offset.
0099	A6		PLO	R6	Put result in R6 as Code pointer.
009A	F805		LDI	05	Put Morse character count in RE.0.
009C	AE		PLO	RE	–
009D	06		LDN	R6	Get Morse pattern via R6.
009E	BE		PHI	RE	Store it temporarily in RE.1.
009F	8E	CHR	GLO	RE	Have five Morse characters
00A0	3288		BZ	CODE	been output yet?
00A2	9E		GHI	RE	If not, get code pattern
00A3	2E		DEC	RE	out of temporary storage in
00A4	76		SHRC		RE.1 and shift character bit
00A5	BE		PHI	RE	into DF.
00A6	33AD		BDF	DASH	If bit = 1 load a dash.
00A8	F819	DOT	LDI	19	Otherwise load a dot.
00AA	A8		PLO	R8	
00AB	3BB0		BNF	EXIT	
00AD	F86B	DASH	LDI	6B	
00AF	A8		PLO	R8	
00B0	F8B6	EXIT	LDI	B6	Restore DOT/DASH sub pointer R7
00B2	A7		PLO	R7	to 00B6 and . . .
00B3	D7		SEP	R7	Gosub DOT/DASH.
00B4	309F		BR	CHR	Back for another character.
00B6	F835	DT/DSH	LDI	35	Load pitch value of tone.
00B8	7B		SEQ		"ON"
00B9	FF01	DECPT1	SMI	01	Decrement pitch value.
00BB	3AB9		BNZ	DECPT1	If "ON" time is not up, dec again.
00BD	F835		LDI	35	If it is, load pitch value again.
00BF	7A		REQ		"OFF"
00C0	FF01	DECPT2	SMI	01	Decrement pitch value.
00C2	3AC0		BNZ	DECPT2	If "OFF" time is not up, dec again.
00C4	28		DEC	R8	
00C5	88		GLO	R8	Has the character been sent yet?
00C6	3AB6		BNZ	DT/DSH	If not, go back for more.
00C8	F80A	SPACE	LDI	0A	Load value for space between
00CA	B8		PHI	R8	characters in R8.1.
00CB	28	DECSP	DEC	R8	Decrement space value.
00CC	98		GHI	R8	Is the space up yet?
00CD	3ACB		BNZ	DECSP	If not, decrement it again.
00CF	D5		SEP	R5	Return to CODE.
00D0	–	BINHRS			Binary Hours stored here.
00D1	–	BINMIN			Binary Minutes stored here.
00D2	–	10'sHR			BCD 10's Hours stored here.
00D3	–	1'sHR			BCD 1's Hours stored here.
00D4	–	10'sMN			BCD 10's Minutes stored here.
00D5	–	1'sMN			BCD 1's Minutes stored here.
00D6	1F	DIGIT TABLE			"0"
00D7	1E				"1"
00D8	1C				"2"
00D9	18				"3"
00DA	10				"4"
00DB	00				"5"
00DC	01				"6"
00DD	03				"7"
00DE	07				"8"
00DF	0F				"9"

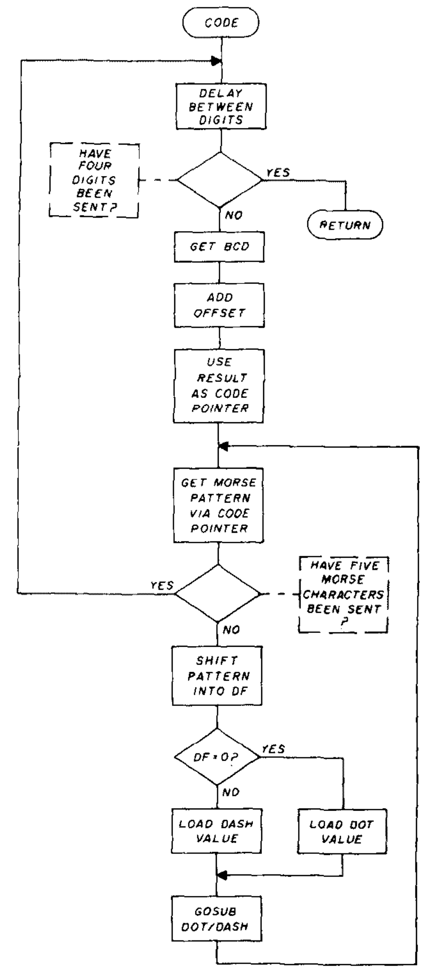


fig. 4. Flowchart of the code routine.

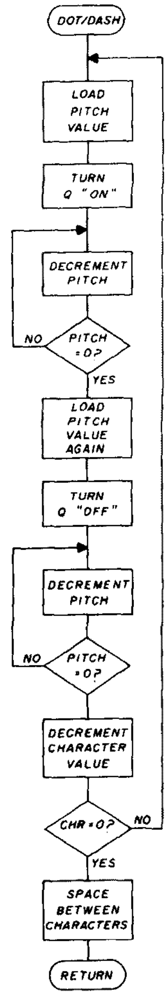


fig. 5. Flowchart of the dot/dash routine.

dot/dash (**fig. 5**). After the code routine fetches a dot or dash, this subroutine generates a tone of proper duration. (A dash is about three times longer than a dot.) The dot/dash routine also generates a space after each character. This space is a period of silence about as long as a dot.

operation and fine-tuning

Before running the program, set the clock by entering the time. To do this, convert the hours and minutes values to hexadecimal. Then enter the hours at location 00D0 and the minutes at location 00D1. Now execute the program from location 0000. The program will give the time whenever the EF4 flag is activated, and automatically on the hour.

Although the program should keep accurate time, you may have to adjust its speed if your microprocessor's clock frequency is different. Do this by varying the value at location 0027. Also, you can tune the pitch of the tone by changing the values at locations 00B7 and 00BE. You can alter the speed of the Morse characters by changing the values for dot, dash, and space at locations 00A9, 00AE, and 00C9. Finally, if you prefer a twelve-hour format, change the value at location 0049 to 0C.

summary

The basic feature of this program is the routine which converts BCD digits into Morse code characters, and in this case, the BCD digits represent time. However, the same routine with some modification could be used where the BCD digits represent something else, like temperature, pressure, voltage, or resistance. You would need more elaborate hardware in these cases, since they involve A/D conversion, but any measuring device could be made to talk with this method.

ham radio

a state-of-the-art
Touchtone® decoder

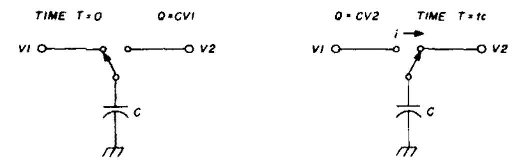
Using Silicon Systems Inc.'s single-chip solution

Silicon Systems inc.'s DTMF (dual-tone, multiple frequency) decoder IC is revolutionizing the way Amateurs use TouchTones®. With this device, it's possible to build a decoder with as few as three ICs, and the resulting circuit (see photo) is small, requires little power, and is very reliable.

a brief history

It wasn't too long ago that every DTMF decoder used and built by Amateurs was made with the NE567 phase-locked loop-tone decoder. At the time, that was the only way to decode dual-tone audio into a useful digital signal; it required tedious adjustment of a potentiometer for every frequency and that adjustment would rarely remain stable when temperature varied.

About five years ago, Mostek released a product that eliminated all the adjustments and made DTMF decoding relatively simple, but rather costly. In the Mostek system, the incoming audio signal is split into the two components of DTMF (i.e., the high-frequency group and the low-frequency group). These two components are then limited and squared before being applied to the Mostek DTMF decoder. Although the cost of the splitting filters is high, this remains a superior system to multiple 567s, as the dynamic range is tremendously improved and no adjustments are necessary.



$$i = \frac{Q(0) - Q(tc)}{tc} = C(V1 - V2)fc$$

$$R = \frac{V}{I} = \frac{(V1 - V2)}{i}$$

$$\rightarrow R = \frac{1}{Cfc}$$

fig. 1. The switched-capacitor principle allows a small capacitor to be used in place of a large resistor. This has allowed the manufacturer to put filters and decoding circuits on one silicon chip.

The next logical step in DTMF decoders was to put the filters, limiters, and squarers on the same chip as the decoder. This was accomplished by Silicon Systems Incorporated (SSI) with their SSI201, a single-chip solution that requires only two small bypass capacitors and a 3.58 MHz color-burst crystal.

operation of the decoder

The major problem was to incorporate rather large capacitors and resistors needed for the filters onto the silicon chip. The largest size capacitor that can be integrated onto a chip is about 100 pF, and even this size requires a large area. Large resistors are not realizable for the same reason. However, a small capacitor can be made to perform, electrically, like a large resistor.

Fig. 1 illustrates the principle of a switched capacitor to realize a large resistor. At time zero, the capacitor is connected to voltage V_1 and the capacitor charges toward the value $Q = CV_1$. At some later time, t_c, the capacitor is switched to voltage V_2 and the value of the charge is $Q = CV_2$. The equations in the figure show the mathematics used to manipulate the values; the last equation is the most interesting: $R = 1/Cf_c$, in other words, a large resistor can be made (electrically) by just using a small capacitor and switching it between voltages at a very fast rate!

This led to the development of the switched-capacitor filtering used in the SSI201 DTMF decoder. (MOS transistors are used as the switches.)

The block diagram of the entire decoder is shown in **fig. 2**. As in the multiple-chip Mostek system, the audio is first split into upper and lower bands. These signals are further filtered to determine the two tones present. Next, the output-decoder circuitry converts this information to digital form, and produces BCD (binary-coded decimal), or optional 2-of-8 outputs. A 3.579545 MHz color-burst crystal is used for the frequency references as well as for the switched-capacitor filter networks.

By Mark Forbes, KC9C, 1000 Shenandoah Drive, Lafayette, Indiana 47905

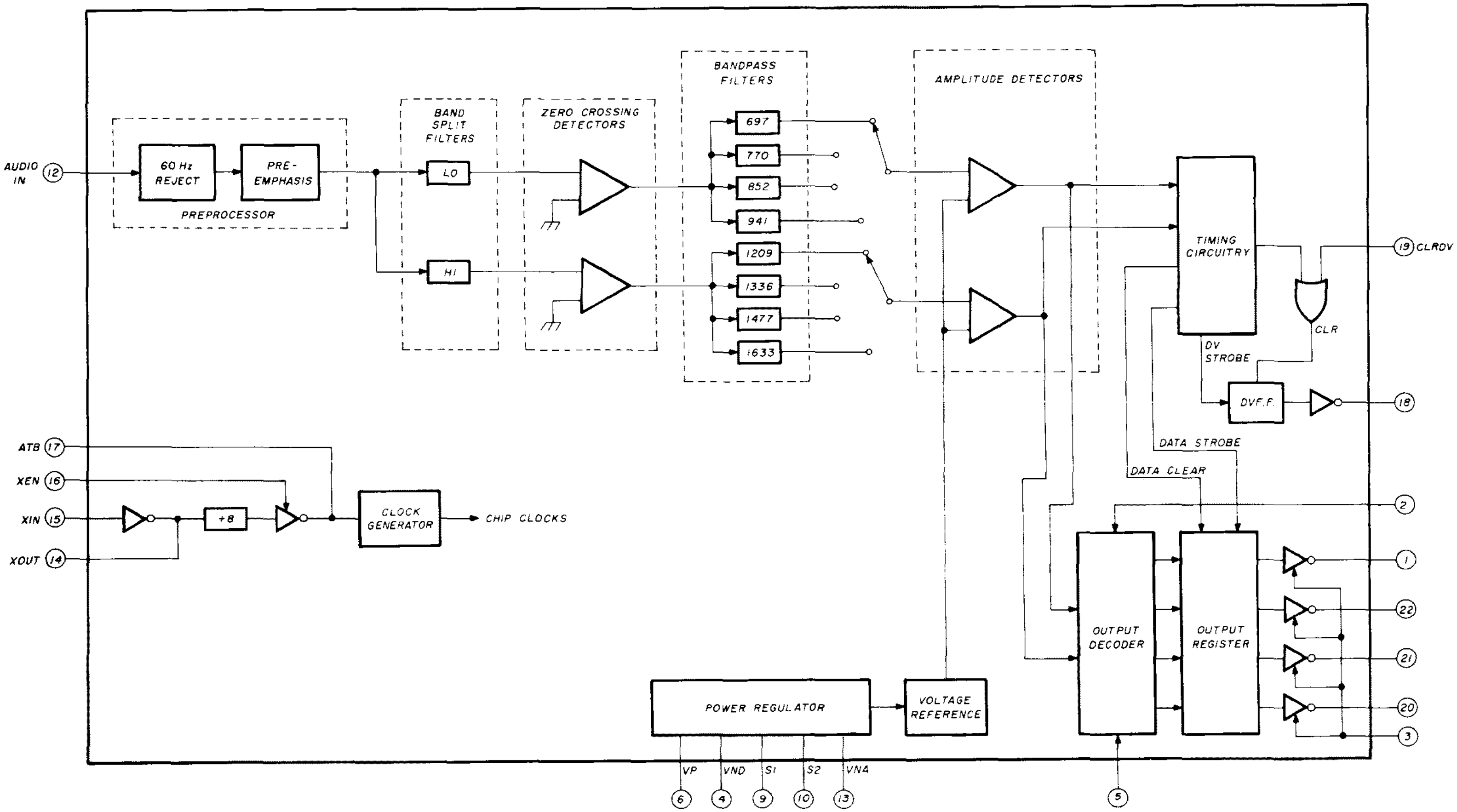


fig. 2. Block diagram of the Silicon Systems, Inc., SSI201 DTMF decoder.

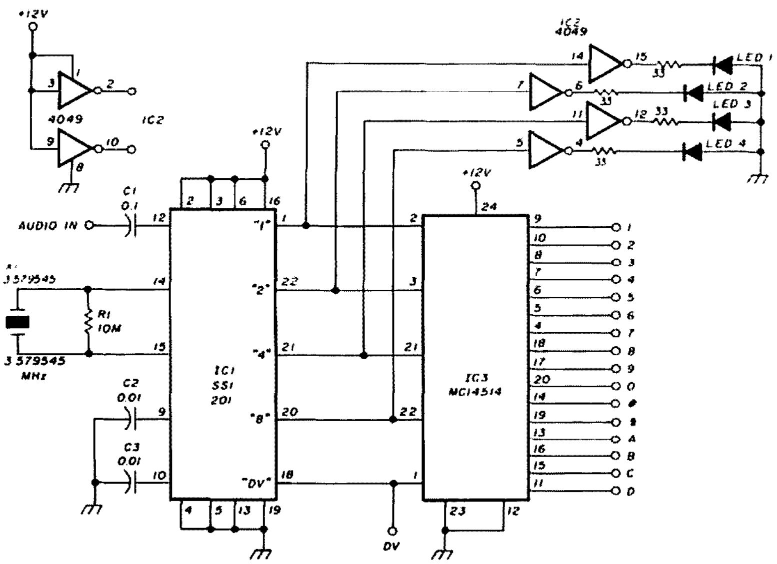


fig. 3. Schematic diagram of the author's completed decoder circuit. The CMOS signals are all 12-volt levels, so a converter is needed if the decoder must drive TTL.

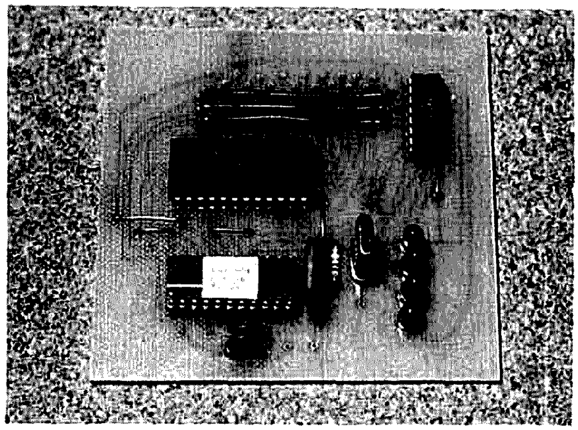
the complete circuit

To make the SSI201 easier to interface to remote-control and repeater circuits, I have added two IC and four LEDs in this DTMF decoder design. The schematic diagram is shown in fig. 3. Audio input is coupled to the SSI201 through a 0.1 μF disk capacitor. The BCD output of the decoder is further decoded into individual "tone-pad" digits by an MC14514B 4-to-16 line demultiplexer.

One useful signal available from the decoder is the DV (data valid) signal. This signal goes high when the output data is in a predefined window of time, and is useful in determining when to sample the outputs of the MC14514 (although these outputs are latched, so the last data remains on the outputs until new data is presented).

As a convenience, LEDs that show the binary value of the decoded output, are included (note: the values for *, Ø, and # are 11, 10, and 12, respectively). A CMOS 4049 inverting buffer is used to drive the LEDs and remove the load from the SSI201.

All the ICs in the project are powered from 12 Vdc. A note of caution here — the SSI201 requires 12 volts and *not 13.8 volts* as found on many power supplies. A small IC voltage regulator will provide the proper 12 volts if you don't have such a power supply (an LM340-12 is one such regulator). If the outputs are to be interfaced to 5-volt logic such as TTL, a voltage converter circuit such as that shown in fig. 4 can be employed.



The DTMF decoder circuit.

Construction of the circuit is very simple, using the printed circuit artwork provided in fig. 5. All that is necessary is to solder the ICs and apply 12 volts. Sockets are recommended to keep the heat of soldering off the ICs and to facilitate replacement should any of the components fail.

applications

The applications of a DTMF decoder seem almost limitless, especially when no adjustments are neces-

sary. The most obvious application is in repeater control. This circuit is highly reliable and not subject to degradation by temperature or variation of signal levels. These features, coupled with the compact size of this circuit, make it perfect for use in repeaters.

A reliable circuit like this one also opens the door to an underexplored facet of Amateur radio: remote control. Remote control of more than just repeaters is allowed by the FCC. In fact, almost anything can be remotely controlled via Amateur radio. Types of applications include remote HF stations, models, or even your house lights.

Another good use of the DTMF decoder is in autopatch circuits. Most autopatches couple the DTMF tones directly to the telephone line from the receiver.

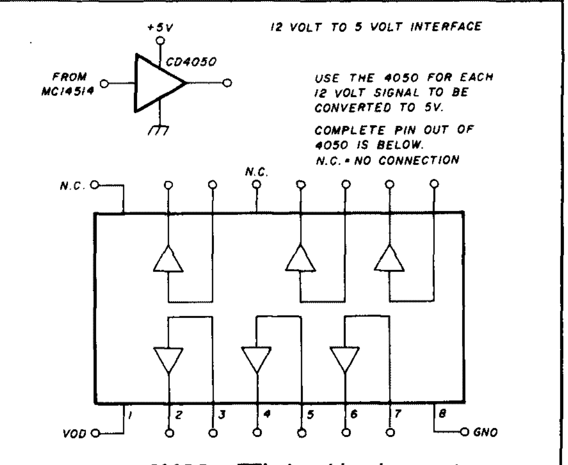


fig. 4. A CMOS to TTL signal-level converter.

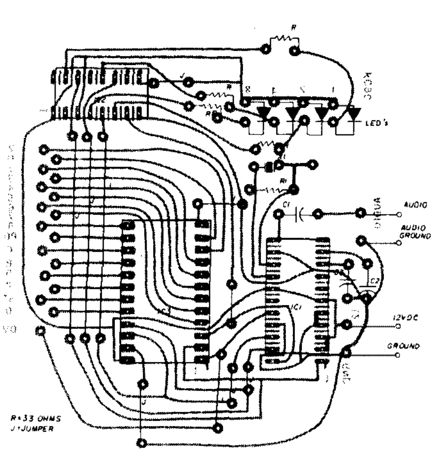


fig. 5. Printed-circuit board artwork

table 1. Parts and Prices List.

part	description	source	price
IC 1	SSI201 DTMF decoder	SAI Marketing*	60.00
IC 2	4049 inverter	Digikey	0.47
IC 3	MC14514 (4514B)	Digikey	1.99
LED 1-4	Light Emitting Diodes	Radio Shack	4 for 1.58
C 1	0.1 uF disk capacitor	Radio Shack	2 for 0.49
C 2,3	0.01 uF disk capacitor	Radio Shack	2 for 0.39
R 1	10 Megohm resistor	Radio Shack	2 for 0.19
X 1	3.579545 MHz crystal	Radio Shack	1.99
PCB	printed circuit board	Author	8.50
		Total Cost	67.10
		Total w/PCB	75.60

Note: Complete parts kits are available from the author for $75.60 plus $1.00 shipping. Or, the ICs and/or PCB may be purchased individually at the listed price plus $1.00 shipping.

*The address of SAI Marketing is: SAI Marketing, Attn. Jim Taylor, 5610 Crawfordsville Road, Indianapolis, Indiana 46224.

This results in two things: the user needing to adjust his TouchTone pad to tight telephone company specifications, and frequently misdialed numbers. By decoding the signal first, then re-encoding with a DTMF generator chip, the telephone line will always have a perfect and precise tone for dialing. And, with the wide dynamic range of the SSI201, adjustment of the user's tone pad is almost never necessary. An additional problem can also be solved: in areas where DTMF dialing is not yet available, a pulse dialer chip in conjunction with the SSI201 can provide autopatch functions.

conclusion

The parts list in **table 1** gives the price and availability of each of the parts at the time of writing. Additionally, I have complete parts kits available for the prices shown, so there should be no trouble in finding all the necessary parts.

The SSI201 is, in my opinion, the best DTMF decoder introduced to date. The Amateur press seems to be behind in the DTMF decoder field. In fact, one book on repeaters published in 1980 still showed 567 circuits for decoding DTMF. The switched capacitor has revolutionized the DTMF scene, and will soon find its way into other areas.

references

Jacobs, G.M., et. al., "Touch-tone decoder chip mates analog filters with digital logic," *Electronics Magazine*, February 15, 1979, McGraw-Hill, Inc.

Silicon Systems Incorporated, "Monolithic Dual-Tone Multi-frequency Receiver Application Note," May 1980.

ham radio

remote control hf operation

An Apple II and Collins KWM-380 talk to each other via the telephone

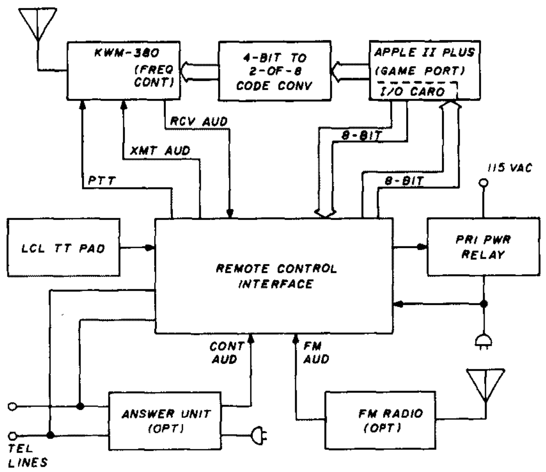


fig. 1. High-frequency remote-control system block diagram.

You can have remote high-frequency radio operation from a TouchTone™ telephone. In this article I explain this design, including the interface used to control the radio and computer; the interface plugs into the radio and computer without modification to either. A remote operator can thereby use a telephone to turn on and off primary power; use a private access code; tune the radio to any discrete frequency or scan up and down; transmit; and have optional fm radio capability. The interface has a safety shutdown feature in case the power or telephone is interrupted.

The remote system is illustrated by the block diagram in **fig. 1**. The center of the system is the interface control, which includes a phone patch, a dual tone multi-frequency (DTMF) decoder, audio amplifiers, and control logic. I use a Rockwell-Collins KWM-380 transceiver with the control interface option, and an Apple II Plus microcomputer with an eight-bit input/output card. A regular phone-answering unit detects the telephone ring. A ring-detection circuit could be incorporated into the interface control, but I prefer having a tape recorder tied to the system for logging. A twelve-button TouchTone™ keypad provides local control. A primary power relay, that includes transient protection, turns on the KWM-380 and the Apple. The phone-answering unit and interface control remain on at all times. An interface device that connects between the Apple's game port and the KWM-380's frequency-control interface connector provides frequency control. An optional fm audio-decoder is also included to provide additional system control and operation from a VHF/UHF fm radio.

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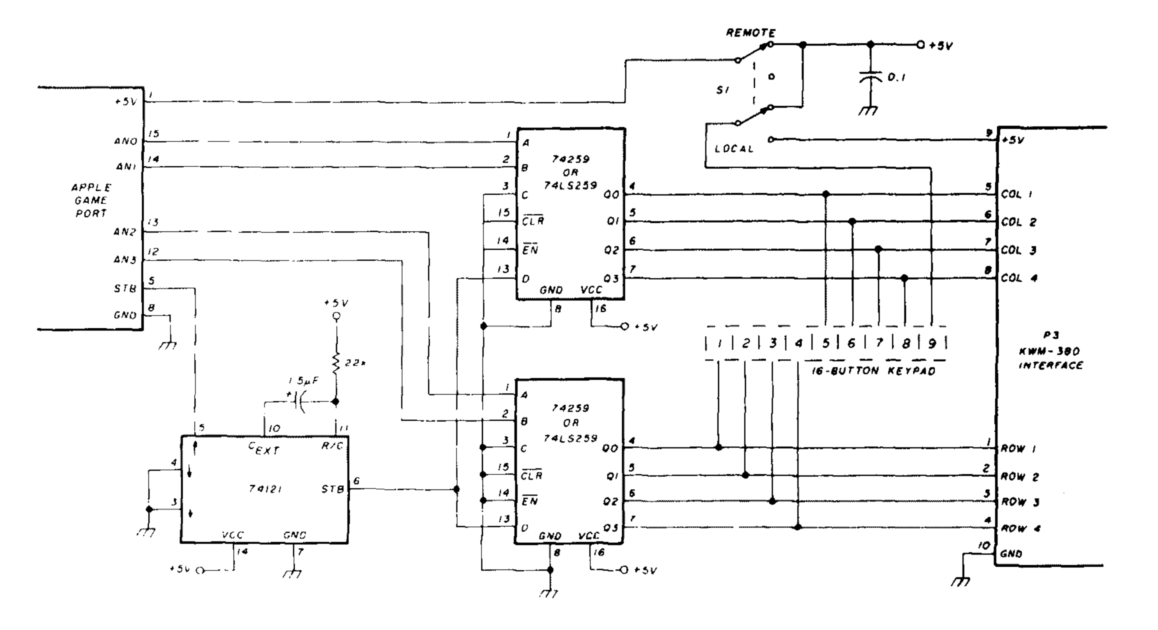


fig. 2. Apple II to KWM-380 interface schematic. It converts a four-bit binary code to a 2-of-8 code with a strobe to load the data into the radio.

The frequency-control circuitry is in a case that contains the KWM-380's sixteen-button keypad. The case also contains a switch that selects +5 volt power from either the radio or the Apple. There are two reasons for interfacing the frequency control separately: the first is that this portion can be a separate project; and the second is that fewer parts are required to build an interface compared to an I/O card to insert into one of the Apple's card slots. The purpose of the interface device is to convert the four binary-outputs and strobe available from the Apple's game port to an eight-bit two-of-eight code required by the KWM-380 (see **fig. 2** for the schematic diagram of the frequency interface). The output of each 74259 decoder is tied directly with the sixteen-button keypad to allow frequency input to the radio while the Apple is running. The negative strobe of the Apple triggers a 74121 one-shot and clocks the data into the radio. If only frequency control from the Apple II is going to be used, lines 2000 through 2650 of the program listing form a routine for operating only frequency control for the KWM-380; delete lines 2030 through 2070 and replace them with a **GET F** statement from the keyboard.

The remote control interface is the heart of the system; **fig. 3** is its functional block diagram and **fig. 4** is its schematic. The phone-answering unit has an earplug that I use to connect the telephone audio to the interface control. After the unit hooks onto the telephone line and sends its outgoing message, it

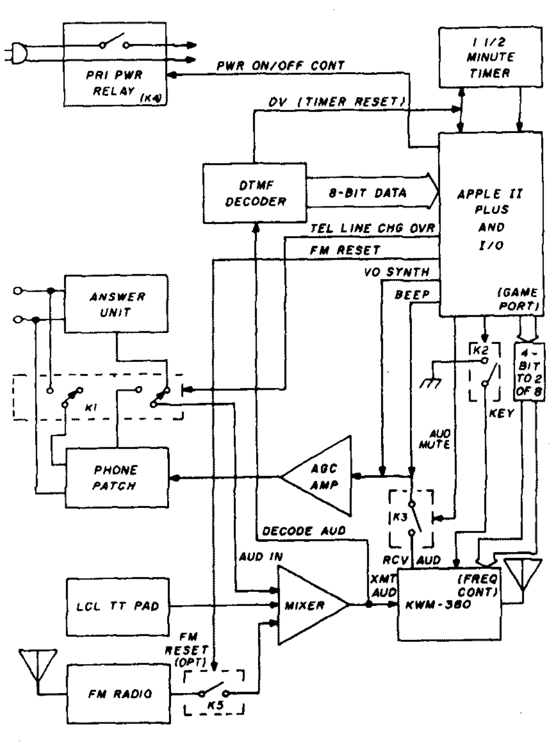


fig. 3. High-frequency remote-control functional block diagram. Each relay function is shown as a dotted line box.

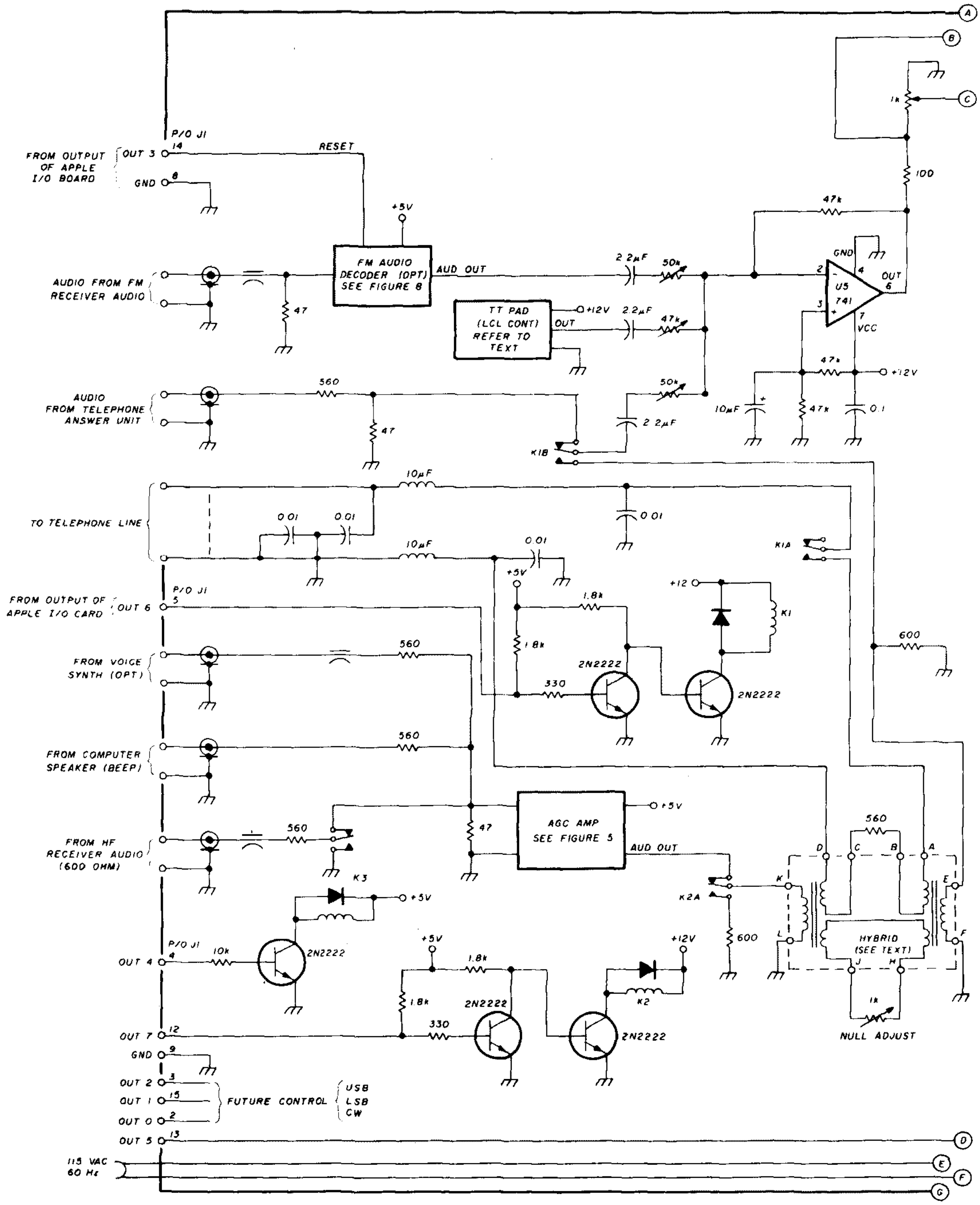
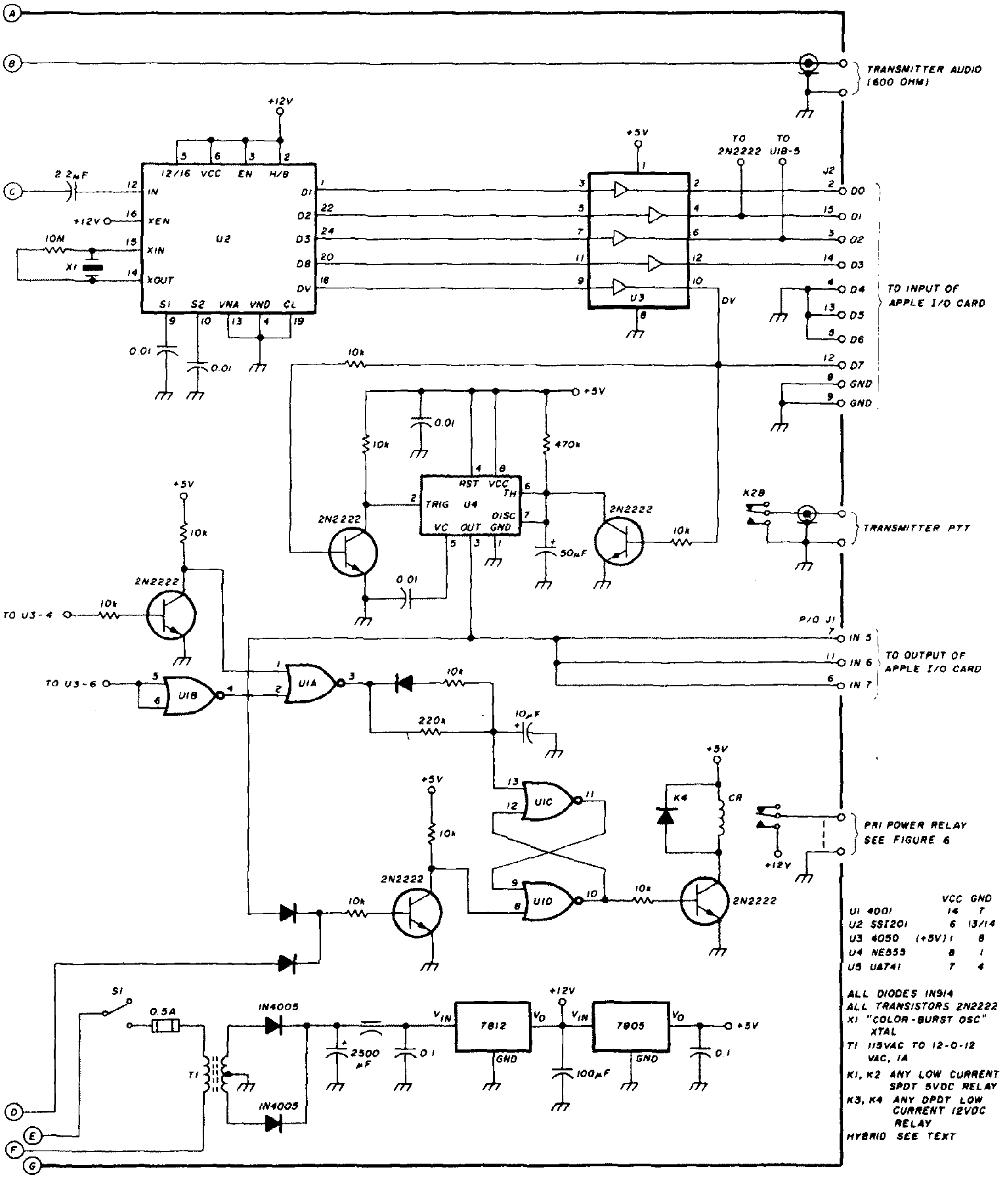


fig. 4. The interface-control schematic diagram. Note that the AGC amplifier and optional fm audio decoder schematic are shown separately.



TRANSMITTER AUDIO (600 OHM)
+12V
+5V
TO 2N2222
TO U1B-5
J2
12/16 VCC EN H/8
IN
XEN
XIN
XOUT
U2
S1 S2 VNA VND CL
D1 D2 D3 D8 DV
2.2µF
10M
X1
0.01
U3
DV
D0 D1 D2 D3 D4 D5 D6 D7 GND GND
TO INPUT OF APPLE I/O CARD
10k
470k
RST VCC
TRIG U4 TH
DISC
VC OUT GND
50µF
2N2222
K2B
TRANSMITTER PTT
TO U3-4
TO U3-6
U1B
U1A
U1C
U1D
P/O J1
IN 5
IN 6
IN 7
TO OUTPUT OF APPLE I/O CARD
220k
10µF
K4
CR
PRI POWER RELAY SEE FIGURE 6
S1
0.5A
T1
1N4005
2500 µF
0.1
VIN
7812
VO
GND
100µF
7805
U1 4001 VCC 14 GND 7
U2 SSI201 VCC 6 GND 13/14
U3 4050 (+5V) 1 8
U4 NE555 8 1
U5 UA741 7 4
ALL DIODES 1N914
ALL TRANSISTORS 2N2222
X1 "COLOR-BURST OSC" XTAL
T1 115VAC TO 12-0-12 VAC, 1A
K1, K2 ANY LOW CURRENT SPDT 5VDC RELAY
K3, K4 ANY DPDT LOW CURRENT 12VDC RELAY
HYBRID SEE TEXT

(fig. 4 cont'd)

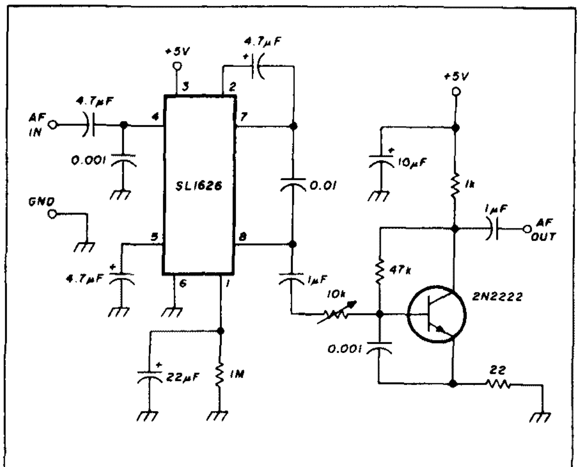


fig. 5. AGC amplifier schematic diagram. This circuit maintains a constant audio level to the phone patch.

allows twenty seconds for an incoming message. During this time you must access the system. The incoming audio is tied through the normally closed contacts of relay K1B directly to U5, a 741 operational-amplifier. U2 is an SSI 201 OTMF CMOS receiver that decodes the incoming audio tones, and U3, a CMOS-to-TTL buffer that passes the data to the eight-bit TTL input of the Apple's I/O card. If the proper access code is present, the output of the Apple pulls in relay K1 and connects the telephone line to the phone patch. The answering unit will drop off by now.

The phone patch contains a transformer-type hybrid with a balancing network. The hybrid transformers that I used were surplus, and no part number is available; the builder must decide upon his own transformers. I'm using a 1-kilohm pot for null adjustment. Some situations may require some series capacitance to null out the telephone line inductance; the system will not work without proper balance. The DTMF decoder requires at least a 12-dB signal-to-noise ratio, which is why a null is important. An AGC amplifier is needed to maintain a constant level to the hybrid. If you were to measure the output of your receiver, you would find the audio level varies by as much as 20 dB. **Fig. 5** is the schematic diagram of the AGC amplifier I'm using. It uses an SL1626 voice-operated gain adjusting device (VOGAD) that drives a simple 2N2222 transistor amplifier. The output is extremely constant and maintains proper audio level. Because the VOGAD operates at low levels, resisting dividers are used to reduce the input to the proper levels. The AGC amplifier controls only outgoing audio, which includes the hf received, the beeps, and possibly a voice synthesizer. Throughout the program, beeps from the Apple's speaker tell the operator where he is during operation. For connecting audio to the system, I couple to the Apple using a 0.47-μF capacitor wrapped between the audio high side of the speaker connector and the interface. For audio low, I connect the grounds together. In the Apple, the speaker is dc-coupled to +5 volts, so be careful when connecting to the Apple's speaker connector (refer to the Apple II reference manual). Incoming audio (tones and voice) from the phone line via the hybrid, the local TT keypad, and optional fm control go to U5, the audio mixer. The output of U5 goes to the DTMF decoder and to the KWM-380 transmit audio.

The control-logic portion of the interface control consists mainly of a timer, a latch, and four control relays. Timer U4, a 555, stays on for one and one-half minutes. It is reset from the data valid (DV) output of U2. If there isn't any key activity before timeout, relays K1 (phone line) and K2 (transmit/receive) drop off. Latch U1, a 4001 quad NOR gate, enables relay K4 (primary power) and turns on the radio and Apple. A shut-off command from data-out 5 causes relay K4 to drop out when U4 times out, and the radio and the Apple will turn off. Relay K3 mutes the high-frequency-received audio when a command from data-out 4 appears. Muting is used when you wish to hear only the beep or voice synthesizer (if used). Relay K2 is the transmit key relay; it sends a ground to the KWM-380's keyline and maintains a 600-ohm load across the input side of the hybrid during transmit. **Table 1** gives a detailed description of each data line and its address (I/O card in slot 4) from BASIC.

Fig. 6 is the schematic for the primary power-relay. It contains varistor transient suppressors and

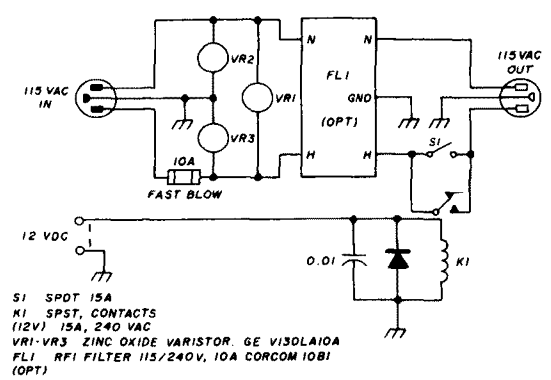


fig. 6. Primary power-relay schematic diagram. This relay box includes additional transient protection (optional). Switch S1 bypasses the relay if remote switching is not desired.

table 1. This table lists the I/O data lines used to interface the KWM-380 and Apple II and gives a description and address from BASIC of each.

data line	description	computer address (slot 4) from BASIC	
input: D0-D7	Parallel eight-bit binary input, MSB is data valid (DV)	PEEK (50176)	
output			
out 7	Pulls in transmit/receive relay K2	POKE 49359,1	on
		POKE 49351,0	off
out 6	Pulls in telephone relay K1	POKE 49358,1	on
		POKE 49350,0	off
out 5	Sets input bit to latch U1C	POKE 49347,0	on
	and U1D to turn off primary	POKE 49355,1	off
	power relay K4		
out 4	Pulls in high-frequency receiver mute relay K3	POKE 49356,1	on
		POKE 49348,0	off
out 3	Resets the fm decoder	POKE 49349,0	
out 2-out 0	Future control to be used to change the KWM-380 mode, between USB, LSB, and CW		
game port:			
AN0		POKE 49241,0	on
		POKE 49240,0	off
AN1	Parallel four-bit binary	POKE 49243,0	on
	code to the two-of-eight	POKE 49242,0	off
AN2	code converter to drive	POKE 49245,0	on
	the KWM-380	POKE 49244,0	off
AN3		POKE 49247,0	on
		POKE 49246,0	off
STROBE	Clocks data into the KWM-380	PEEK (-16320)	

an EMI filter. These aren't necessary, but I had them in my junk box, so I used them. Power is switched on when K4 supplies +12 volts to relay K1, located in the primary power-relay box. When the system is on and I'm away from home, I feel secure knowing there is some protection. Not shown is a 115-Vac antenna change-over relay that grounds the input to the receiver when power is off; when power is on, the antenna is ungrounded. The power supply uses 7812 and 7805 voltage regulators. The entire interface control operates from +12 volts and +5 volts. Fig. 7 is the diagram of the interconnection between the interface control and the Apple's eight-bit I/O card. An optional goodie is the fm audio-decoder, whose schematic is shown in fig. 8. It allows direct access to the computer through the DTMF decoder via fm radio. This is used in case you want to operate remotely from VHF or UHF. The tone decoders are 567s and can be adjusted to detect any dual tone; I'm using tones from my keypad. It is activated by holding the proper key for eight seconds; both the telephone and fm radio operate the system, or the fm radio can operate alone. A command from data-out 3 resets the decoder (turns the fm audio off).

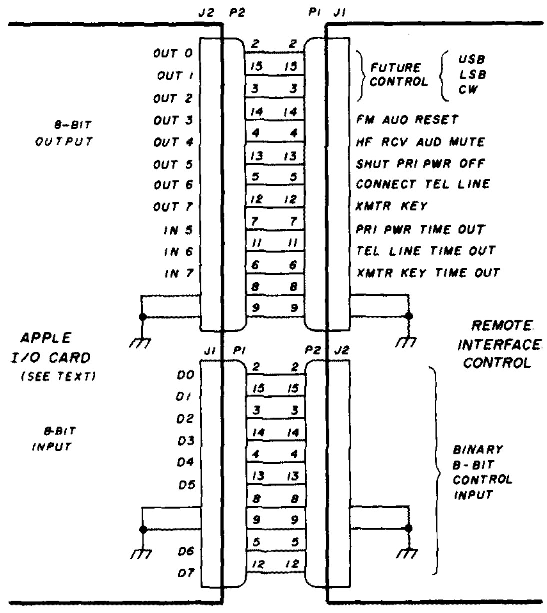


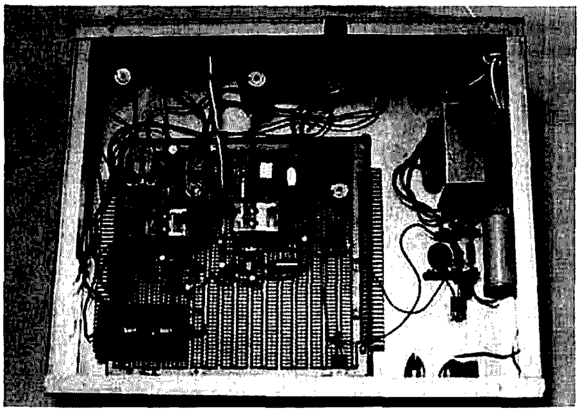
fig. 7. Wiring diagram from the Apple II to the remote interface control.

system operation

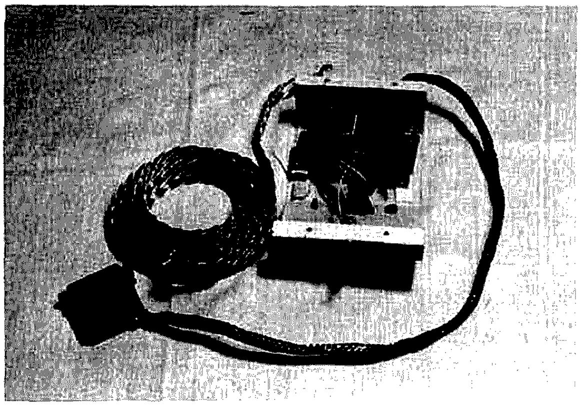
For testing, replace the telephone line with a 900-ohm resistor to provide balance to the hybrid. **Fig. 9** is the BASIC program. The program as listed will not autoboot; after the program is typed in and saved, insert a new disk and type: **INIT HELLO**. Apple DOS will create an autoboot disk. If the radio and Apple are off, push the digit 6 on the local TT keypad for five seconds. This allows U1A to charge the 10-μF delay capacitor to set latch U1C and U1D and enable relay K4. System power will now be on. Line 70 is the three-digit access code; this can be changed at will. I use 789 in this program.

Enter the access code and the program menu, which give prompts to each of the functional subroutines that will appear. This portion of the program is lines 400 through 540. There are six

This picture shows all the components that compose the high-frequency remote-control system. See fig. 1 for the block diagram.



This picture shows the remote control interface. This unit contains a phone patch, a DTMF decoder, level amplifiers, control logic and relays. See fig. 4 for its schematic. Note that space is available on the circuit board to fully remote the KWM-380.



This picture shows the frequency interface. It connects between the KWM-380 control interface connector and the Apple game port. See fig. 2 for its schematic.

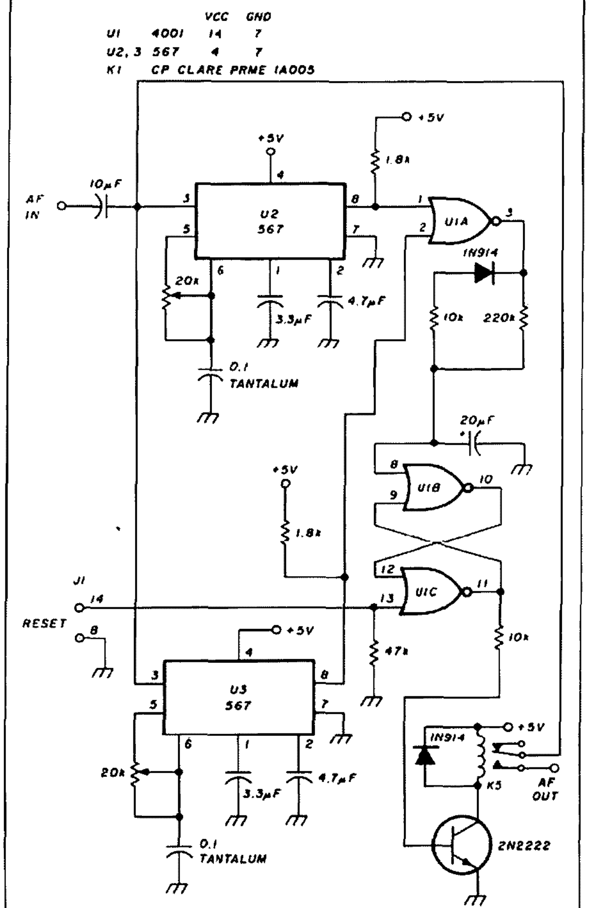


fig. 8. Optional fm audio-decoder schematic diagram. Any pair of tones can be selected. The tones must be held on for about eight seconds before relay K1 pulls in. This permits the system to be operated by an fm radio or telephone link.

subroutines, each of which can jump to its particular function when called. Lines 7000 through 7050 show how the Apple gets incoming data that is not from the keyboard.

For the Mute subroutine, enter 1; one beep sounds. This allows the # key to *silence* the KWM-380 or * key to return the audio. The subroutine will automatically return to the menu.

For the Frequency Enter subroutine, enter 2; two beeps sound. This subroutine allows the operator to enter any frequency. A * is used for the decimal place and # loads the KWM-380 and returns to the menu. An example for entering 14.225 MHz is **14*225#**.

For the Scan subroutine, enter 3; three beeps sound. Entering 1 makes the radio scan up. Entering 2 stops the radio from scanning. Entering 3 starts the radio scanning down. Entering * bumps the radio up *1 kHz; 7 bumps it up 100 Hz. Entering # bumps the* radio down 1 kHz; **9** bumps it down 100 Hz. To return back to the menu, enter **0**.

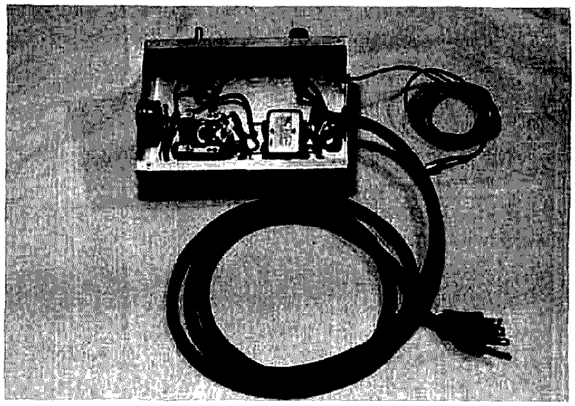
For the Control Option subroutine, enter **4**; four beeps sound. The Control Option subroutine allows # to reset the *fm radio or* * to shut off the primary power after you exit the program. This subroutine automatically returns to the menu.

fig. 9. BASIC program listing.

```
10  HOME : VTAB 12
20  PRINT "  <<<<<    REMOTE CONTROL    >>>>>"
30  HTAB 10: PRINT "WRITTEN BY DICK SANDER"
35  POKE 49358,1
40  FOR O = 1 TO 50: NEXT O
45  CALL  - 198
50  POKE 49350,0
52  HOME
54  REM  UNKEY:POKE 49357,1
56  POKE 49356,1: POKE 49347,0
60  PRINT "INPUT ACCESS CODE (3 DIGITS)"
70 AX = 7:AY = 8:AZ = 9
75  PRINT : PRINT "ACCESS CODE IS:": PRINT : PRINT
80  PRINT AX,AY,AZ
90  GOSUB 7000
100  IF B < AX OR B > AX THEN  GOTO 90
110  GOSUB 7000
120  IF B < AY OR B > AY THEN  GOTO 90
130  GOSUB 7000
140  IF B < AZ OR B > A2 THEN  GOTO 90
150  POKE 49358,1
160  POKE 49357,1
170  REM   MUTE:POKE 49356,1
180  REM  PWR ON:POKE 49347,0
400  REM  MENU
410  HOME
420  VTAB 3: HTAB 18: PRINT "MENU": PRINT
430  PRINT : HTAB 10
440  PRINT "1. ENABLE RECEIVER"
450  PRINT : HTAB 10
460  PRINT "2. ENTER FREQUENCY"
470  PRINT : HTAB 10
480  PRINT "3. SCAN FREQUENCY"
490  PRINT : HTAB 10
500  PRINT "4. CONTROL OPTIONS"
510  PRINT : HTAB 10
520  PRINT "5. TRANSMIT"
530  PRINT : HTAB 10
540  PRINT "6. EXIT"
690 A =  PEEK (50176)
700  IF A < 128 THEN 690
710 A = A - 128
720  IF A = 10 THEN A = 0
730  IF  PEEK (50176) > 127 THEN 730
740  IF A < 1 OR A > 6 THEN  CALL  - 198: GOTO 690
750  IF A = 1 THEN  GOTO 1000
760  IF A = 2 THEN  GOTO 2000
770  IF A = 3 THEN  GOTO 3000
780  IF A = 4 THEN  GOTO 4000
790  IF A = 5 THEN  GOTO 5000
800  IF A = 6 THEN  GOTO 6000
1000  REM   ENABLE RCVR
1010  CALL  - 198
1020  HOME : PRINT "* ENABLES RCVR - # DISABLES RCVR "
1030  GOSUB 7000
1040  IF B = 11 THEN  POKE 49356,1: CALL  - 198: GOTO 410
1050  IF B = 12 THEN  POKE 49348,0: GOTO 410
1060  IF B > 12 THEN  GOTO 1000
1070  IF B < 11 THEN  GOTO 1000
2000  REM  INPUT FREQUENCY
2010  CALL  - 198: CALL  - 198
2020  HOME : PRINT "ENTER FREBUENCY"
2025 SF$ = ""
2030 F =  PEEK (50176)
2040  IF F < 128 THEN 2030
2050 F = F - 128
2060  IF F = 10 THEN F = 0
2070  IF  PEEK (50176) > 127 THEN 2070
2080  IF F = 0 THEN  GOSUB 2250:SF$ = SF$ + "0"
2090  IF F = 1 THEN  GOSUB 2280:SF$ = SF$ + "1"
2100  IF F = 2 THEN  GOSUB 2310:SF$ = SF$ + "2"
2110  IF F = 3 THEN  GOSUB 2340:SF$ = SF$ + "3"
2120  IF F = 4 THEN  GOSUB 2370:SF$ = SF$ + "4"
2130  IF F = 5 THEN  GOSUB 2400:SF$ = SF$ + "5"
2140  IF F = 6 THEN  GOSUB 2430:SF$ = SF$ + "6"
2150  IF F = 7 THEN  GOSUB 2460:SF$ = SF$ + "7"
2160  IF F = 8 THEN  GOSUB 2490:SF$ = SF$ + "8"
2170  IF F = 9 THEN  GOSUB 2520:SF$ = SF$ + "9"
2180  IF F = 11 THEN  GOSUB 2580:SF$ = SF$ + "."
2190  IF F = 12 THEN  GOTO 2220
2200  PRINT F
2210  GOTO 2030
2220  GOSUB 2550
2230  GOTO 400
```

```
2240 GOTO 2030
2250 POKE 49241,0: POKE 49242,0: POKE 49245,0: POKE 49247,0
2260 GOSUB 2610
2270 RETURN
2280 POKE 49240,0: POKE 49242,0: POKE 49244,0: POKE 49246,0
2290 GOSUB 2610
2300 RETURN
2310 POKE 49241,0: POKE 49242,0: POKE 49244,0: POKE 49246,0
2320 GOSUB 2610
2330 RETURN
2340 POKE 49240,0: POKE 49243,0: POKE 49244,0: POKE 49246,0
2350 GOSUB 2610
2360 RETURN
2370 POKE 49240,0: POKE 49242,0: POKE 49245,0: POKE 49246,0
2380 GOSUB 2610
2390 RETURN
2400 POKE 49241,0: POKE 49242,0: POKE 49245,0: POKE 49246,0
2410 GOSUB 2610
2420 RETURN
2430 POKE 49240,0: POKE 49243,0: POKE 49245,0: POKE 49246,0
2440 GOSUB 2610
2450 RETURN
2460 POKE 49240,0: POKE 49242,0: POKE 49244,0: POKE 49247,0
2470 GOSUB 2610
2480 RETURN
2490 POKE 49241,0: POKE 49242,0: POKE 49244,0: POKE 49247,0
2500 GOSUB 2610
2510 RETURN
2520 POKE 49240,0: POKE 49243,0: POKE 49244,0: POKE 49247,0
2530 GOSUB 2610
2540 RETURN
2550 POKE 49240,0: POKE 49243,0: POKE 49245,0: POKE 49247,0
2560 GOSUB 2610
2570 RETURN
2580 POKE 49240,0: POKE 49242,0: POKE 49245,0: POKE 49247,0
2590 GOSUB 2610
2600 RETURN
2610 REM STROBE ROUTINE
2620 ZZ = PEEK ( - 16320)
2650 RETURN
3000 REM SCAN FREQ
3125 CALL - 198: CALL - 198: CALL - 198
3130 HOME : PRINT "STARTING FREQUENCY? "
3135 PRINT : PRINT
3140 HOME : VTAB 24: HTAB 20: PRINT SF$
3145 VTAB 5: HTAB 10
3150 PRINT "PUSH 1 TO INCREASE FREQUENCY"
3155 VTAB 7: HTAB 10
3160 PRINT "PUSH 2 TO STOP SCANNING"
3165 VTAB 9: HTAB 10
3170 PRINT "PUSH 3 TO DECREASE FREQUENCY"
3175 PRINT : PRINT
3180 T = PEEK (50176)
3185 IF T < 128 THEN 3180
3190 T = T - 128
3195 IF T = 10 THEN T = 0
3200 IF T = 1 THEN GOTO 3225
3205 IF T = 2 THEN GOTO 3180
3210 IF T = 3 THEN GOTO 3285
3211 IF T = 7 THEN GOTO 3750
3212 IF T = 9 THEN GOTO 3750
3213 IF T = 11 THEN GOTO 3750
3214 IF T = 12 THEN GOTO 3750
3215 IF T = 0 THEN CALL - 198: GOTO 410
3220 IF T > 3 THEN CALL - 198: GOTO 3180
3225 REM SCAN UP
3230 X = VAL (SF$)
3235 Y = X + 100
3240 FOR U = X TO Y STEP .00030
3245 REM CHECK FOR NEW KEY
3250 IF PEEK (50176) < 128 THEN 3260
3255 IF PEEK (50176) - 128 = 2 THEN HOME : CALL - 198: GOTO 3140
3260 VTAB 20: HTAB 10
3265 PRINT U;J = U: GOSUB 3695
3270 SF$ = STR$ (U)
3275 NEXT U
3280 GOTO 3140
3285 REM SCAN DOWN
3290 X = VAL (SF$)
3295 Y = X - 100
3300 FOR DN = X TO Y STEP - .00030
3305 REM CHECK FOR STOP KEY
3310 IF PEEK (50176) < 128 THEN GOTO 3320
3315 IF PEEK (50176) - 128 = 2 THEN HOME : CALL - 198: GOTO 3140
3320 VTAB 20: HTAB 10
3325 PRINT DN;J = DN: GOSUB 3695
3330 SF$ = STR$ (DN)
3335 NEXT DN
3340 GOTO 3140
3345 FOR I = 1 TO 9
3350 IF I = 1 THEN IF LEFT$ (C$,1) = "." THEN B$ = "0": GOSUB 3375
3355 B$ = MID$ (C$,I,1): GOSUB 3375
3360 IF B$ = "E" THEN 3370
3365 NEXT I
3370 RETURN
3375 K = VAL (B$) + 1
3380 IF K < > 1 THEN 3395
3385 IF B$ = "0" THEN 3395
3390 GOTO 3405
3395 ON K GOSUB 3440,3455,3470,3485,3500,3515,3530,3545,3560,3575
3400 RETURN
3405 IF B$ = "A" THEN GOSUB 3590
3410 IF B$ = "B" THEN GOSUB 3605
3415 IF B$ = "C" THEN GOSUB 3620
3420 IF B$ = "D" THEN GOSUB 3635
3425 IF B$ = "E" THEN GOSUB 3650
3430 IF B$ = "." THEN GOSUB 3665
3435 RETURN
3440 POKE 49241,0: POKE 49242,0: POKE 49245,0: POKE 49247,0
3445 GOSUB 3680
3450 RETURN
3455 POKE 49240,0: POKE 49242,0: POKE 49244,0: POKE 49246,0
```

```
3460  GOSUB 3680
3465  RETURN
3470  POKE 49241,0: POKE 49242,0: POKE 49244,0: POKE 49246,0
3475  GOSUB 3680
3480  RETURN
3485  POKE 49240,0: POKE 49243,0: POKE 49244,0: POKE 49246,0
3490  GOSUB 3680
3495  RETURN
3500  POKE 49240,0: POKE 49242,0: POKE 49245,0: POKE 49246,0
3505  GOSUB 3680
3510  RETURN
3515  POKE 49241,0: POKE 49242,0: POKE 49245,0: POKE 49246,0
3520  GOSUB 3680
3525  RETURN
3530  POKE 49240,0: POKE 49243,0: POKE 49245,0: POKE 49246,0
3535  GOSUB 3680
3540  RETURN
3545  POKE 49240,0: POKE 49242,0: POKE 49244,0: POKE 49247,0
3550  GOSUB 3680
3555  RETURN
3560  POKE 49241,0: POKE 49242,0: POKE 49244,0: POKE 49247,0
3565  GOSUB 3680
3570  RETURN
3575  POKE 49240,0: POKE 49243,0: POKE 49244,0: POKE 49247,0
3580  GOSUB 3680
3585  RETURN
3590  POKE 49241,0: POKE 49243,0: POKE 49244,0: POKE 49246,0
3595  GOSUB 3680
3600  RETURN
3605  POKE 49241,0: POKE 49243,0: POKE 49245,0: POKE 49246,0
3610  GOSUB 3680
3615  RETURN
3620  POKE 49241,0: POKE 49243,0: POKE 49244,0: POKE 49247,0
3625  GOSUB 3680
3630  RETURN
3635  POKE 49241,0: POKE 49243,0: POKE 49245,0: POKE 49247,0
3640  GOSUB 3680
3645  RETURN
3650  POKE 49240,0: POKE 49243,0: POKE 49245,0: POKE 49247,0
3655  GOSUB 3680
3660  RETURN
3665  POKE 49240,0: POKE 49242,0: POKE 49245,0: POKE 49247,0
3670  GOSUB 3680
3675  RETURN
3680  REM  STROBE ROUTINE
3685 ZZ =  PEEK ( - 16320)
3690  RETURN
3695 C$ =  STR$ (J) + "000000000"
3700  IF J < 1 THEN X = 6: GOTO 3715
3705  IF J < 10 THEN X = 7: GOTO 3715
3710 X = 8
3715 C$ =  LEFT$ (C$,X) + "E"
3720  GOSUB 3345
3725  RETURN
3750  REM  BUMP FREQ UP OR DOWN
3755 ST =  VAL (SF$)
3760  IF T = 11 THEN DL = .001
3765  IF T = 12 THEN DL =  - .001
3770  IF T = 7 THEN DL = .0001
3780 J = ST + DL
3785 SF$ =  STR$ (J)
3790  GOSUB 3695
3795  GOTO 3140
4000  REM  CONTROL OPTIONS
4010  CALL  - 198: CALL  - 198: CALL  - 198: CALL  - 198
4020  HOME
4025  PRINT "CONTROL OPTIONS:": PRINT : PRINT
4027  PRINT "   ENTER $ TO MUTE FM RADIO"
4028  PRINT "   ENTER $ TO SHUT OFF POWER"
4030  GOSUB 7000
4105  IF B = 11 THEN  GOTO 4110
4107  IF B = 12 THEN  GOTO 4140
4109  IF B < 11 THEN  GOTO 400
4110  POKE 49349,0
4120  FOR D = 1 TO 10: NEXT D
4130  POKE 49357,1
4135  CALL  - 198: GOTO 400
4140  POKE 49355,1
4170  CALL  - 198: GOTO 400
5000  REM  ENABLE XMTR
5010  CALL  - 198: CALL  - 198: CALL  - 198: CALL  - 198: CALL  - 19
5020  HOME : PRINT "$ KEYS XMTR - $ UNKEYS XMTR"
5030  GOSUB 7000
5040  IF B < 11 THEN  POKE 49351,0: CALL  - 198: GOTO 410
5050  IF B = 11 THEN  GOSUB 5090
5060  IF B = 12 THEN  GOSUB 5120
5070  IF B > 12 THEN  GOTO 410
5080  IF B = 11 THEN 5020
5085  IF B = 12 THEN 5020
5090  HOME
5100  POKE 49359,1
5110  RETURN
5120  HOME
5130  POKE 49351,0
5140  RETURN
6000  REM  EXIT
6010  REM  DISCONNECT RCVR & PHONE
6020  POKE 49350,0
6030  REM  UNKEY XMTR
6040  POKE 49351,0
6050  REM   10 MHZ WWV
6060  GOSUB 2280
6070  GOSUB 2250
6080  GOSUB 2580
6090  GOSUB 2550
6100  GOTO 75
7000 B =  PEEK (50176)
7010  IF B < 128 THEN 7000
7020 B = B - 128
7030  IF B = 10 THEN B = 0
7040  IF  PEEK (50176) > 127 THEN 7040
7050  RETURN
```



This picture shows the primary power relay. Note the transient protection varistors and rfi filter. Relay K4, located inside the remote interface, controls power relay K1. See fig. 6 for its schematic.

For the Transmit subroutine, enter **5**; five beeps sound. Entering * keys the transmitter; # unkeys the transmitter. Any *digit* will return to the menu.

For the Exit subroutine, enter **6**; the interface control will disconnect the telephone line, tune the radio to WWV, and wait for another call; and turn off power if the control option sets power to off.

There are several smaller projects within this project. I have just touched on each, but I feel there is enough information here to reconstruct my system. The program listing does not contain any voice synthesizer coding; my system does, and it also contains the proper card. I use the voice talker to echo back the frequency after I've entered it, or when I stop scanning.

The system described in this article works reliably without a voice synthesizer. The KWM-380's remote interface allows frequency control only; so for now, I only operate 10/15/20 meters USB with the antenna connected to my tri-band beam. Fortunately, the engineers at Collins left the door open for full remote-control.

I'd like to mention what the future holds for this system. I will add mode selection for the KWM-380, to switch USB, LSB, or CW, along with the proper filter and passband tuning. Also, as an addition to this system or as a stand-alone project for the Apple, I will have an interface to my rotator for beam-heading control.

I really enjoy operating during my breaks at work; so far I've worked about twenty countries remotely. I've found one DX-pedition by using the scan mode.

I would also like to thank Tom McDermott, N5EG, for his technical assistance in this project.

ham radio

technical forum

Welcome to the *ham radio* Technical Forum. The purpose of this feature is to help you, the reader, find answers to your questions, and to give you a chance to answer the questions of your fellow Radio Amateurs. Do you have a question? Send it in!

diesel generator repair

Our organization has a government-surplus 10-kW diesel generator in need of repair. The battery recharging circuitry has been completely destroyed. The unit bears the following markings and information.

Unit markings:
Fermont Engine Generator plant
Division Dynamics Corp. of America
Bridgeport, Conn.
Model # J-141-1 Contract # J-141
Serial # J-141-0018
10 kW 12.5 KVA
PF .80 120/208 V 35 A
60 cy. 3 phase 1800 RPM
Temp. rise 70°
Generator markings:
General Electric # LC7470B16 Type 6J
Model # 5SJ4254P22Y12 Figure 2 generator
Dia/cen. 2261 Frame 254Y
Damaged unit markings:
Fermont # 6064-0001

Please contact us if you have a unit like this. We are in great need of any schematics, manuals, or other information on this unit, and will gladly make arrangements to obtain copies of this information.

The Division of Disaster and Emergency Service is a volunteer search-and-rescue group. We would greatly appreciate any assistance that can be supplied by the readers of *ham radio*.

Wayne Richardson
Lebanon Junction Area Coord.
Bullitt Co. Div. Disaster & Emergency Services
Main Street
Lebanon Junction, KY 40150

another 10-meter beacon

I am writing to inform you that I have designed and built a beacon controller and transmitter and that it is currently in (what I hope will be) permanent operation on 28.208 MHz. The beacon runs twenty-four hours a day, seven days a week, with an input power of 75 watts CW. QSL information is transmitted along with the beacon transmission.

I hope that *ham radio* readers will find this a propagation aid; and the presence of this signal should indicate when the band is open into New England. The antenna is a ground plane at a height of 20 feet (6.1 meters) with 16 one-wavelength long radials.

(I am presently looking for donations of old Novice transmitters which might make a suitable replacement for my current transmitter, should the need arise. Keeping a transmitter on the air continuously can be quite taxing to transmitters designed for Amateur use. I would particularly like to find a Drake 2NT or a DX-60A.)

Leonard J. Umina, WA1IOB
607 Sudbury Street
Marlboro, MA 01752

I am considering transistorizing my old Drake TR-3 transceiver. I do not wish to build or buy the plug-in units that operate from the 250-volt supply in the TR-3. I propose to rectify and filter the 12.6-volt ac originally used for the heaters.

The TR-3's i-f stages use 12BA6 tubes, with plate resistance of 1 megohm and transconductance of 4400 micromhos. I haven't found any single transistor which will match these characteristics, along with high input impedance. Of course I would like to use a single transistor, but I am willing to use two per stage if necessary. Can you help? — Farrell A. Buckley, AK7N

One solution to your question is to use the Solid State Tubes sold by Sartori Associates, P.O. Box 2085, Richardson, Texas 75080. They offer a replacement for the 12BA6. Other solutions are no doubt possible. Perhaps one of our readers can offer a suggestion?

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

More and more Amateurs are faced with the problem of getting on the air from a location where a full-size antenna cannot be erected. What's the answer? Stay on 2 meters and work the local repeater? If only the high-frequency antenna could be magically reduced in size!

Mini-antennas have been used on the high-frequency bands for a long time, the most compact type being the loaded whips for mobile service. While these ultra-short antennas do work, their efficiency is very low (of the order of one or two percent) and their bandwidth is very restricted. As the antenna shrinks in size, compared to the length of the radio wave, efficiency drops and bandwidth decreases. However, it is possible to strike a compromise and achieve good efficiency in an antenna that is smaller than the classic half-wave dipole.

the loaded antenna

Serious investigation of the coil-loaded short antenna started about 1933 when the General Electric Company developed experimental radios for the new mobile police communications system working on the "ultra-high" frequency of about 35 MHz. A summary of the results appeared in the September, 1934, issue of *QST*. The investigation was continued in 1940 by the National Park Service. The N.P.S. wanted 2-4 MHz mobile operation for the mountainous regions of the National parks, many of which exhibit VHF blind spots.[2]

```
100 REM PROGRAM FROM CQ MAG DEC 1981 BY DICK BANDER,  K5QY
200 REM PUT ON TRS-80 AND EXPANDED BY W6EDE FOR W6SAI
400 CLS
500 PRINT"          **   LOADED DIPOLE DESIGN PROGRAM   **"
510 PRINT
520 PRINT "THIS PROGRAM REDUCES ANTENNA TO 1/2 ACTUAL SIZE AND"
525 PRINT
530 PRINT "INSERTS THE COIL IN THE CENTER OF EACH SIDE"
600 PRINT
700 INPUT"HOW MANY FREQ ARE YOU GOING TO INPUT (10 MAX)";Z
710 FOR I=1 TO Z
720 PRINT "FREQ #";I;: INPUT F(I)
730 NEXT
740 FOR I=1 TO Z
750 F=F(I)
900 A=234/F
1100 B=58.5/F
1300 X=((234/F)-B)
1400 Y=((A/2)-B)
1450 IF D>0 THEN 2100
1500 INPUT"DO YOU WANT TO USE THE WIRE TABLE ( Y OR N )";Y$
1600 IF Y$="Y" THEN GOSUB 3000
1700 IF Y$="N" THEN GOSUB 4300
1800 IF Y$="Y" THEN 2100
1900 IF Y$="N" THEN 2100
2000 GOTO 1500
2100 S1=1.E+6/(68*3.1415↑2*F↑2)
2200 S2=1/X*(LOG(24*X/D)-1)
2300 S3=(1-F*B/234)↑2-1
2400 S4=1/Y*(LOG(24*Y/D)-1)
2500 S5=(Y*F/234)↑2-1
2600 S6=S1*(S2*S3-S4*S5)
2650 IF I=1 THEN 2700
2660 GOTO 2703
2700 LPRINT" FREQ","TOTAL","CENTER","COIL","WIRE"
2701 LPRINT" MHZ","LENGHT","TO","IND","DIAM."
2702 LPRINT,"FEET","COIL","UHY","INCHES"
2703 LPRINT
2705 LPRINT F,A,B,S6,D
2900 NEXT I
2901 END
3000 PRINT
3100 INPUT "WIRE GAUGE (#10 TO #22)      ";D
3200 IF D=10 THEN D=.101:RETURN
3300 IF D=12 THEN D=.081:RETURN
3400 IF D=14 THEN D=.064:RETURN
3500 IF D=16 THEN D=.058:RETURN
3600 IF D=18 THEN D=.040:RETURN
3700 IF D=20 THEN D=.032:RETURN
3800 IF D=22 THEN D=.025:RETURN
3900 PRINT
4000 PRINT"WIRE GAUGES MUST BE AN EVEN"
4100 PRINT"NUMBER IN THE RANGE 10-22"
4200 PRINT:PRINT: GOTO 3000
4300 PRINT
4400 INPUT"WHAT IS THE ELEMENT DIAMETER IN INCHES";D
4500 RETURN
```

fig. 1. Loaded dipole program for the TRS-80.

The conclusions of both these investigations point up that a very short, loaded antenna could be made to work well provided it was properly designed. One of the main requirements of proper design was that a high-Q loading coil be used, and that it be placed near the center of the antenna section.

It was there that the matter rested until Jerry Hall, K1PLP, published a classic article in the September, 1974, issue of *QST*, giving a procedure for determining the inductance of a loading coil no matter where it was placed in an antenna.[3] Jerry's example used a dipole instead of a mobile whip. This interesting mathematical exercise was converted into a computer program by Dick Sander, K5-QY, and published in the December, 1981, issue of *CQ*.[4] The short, loaded antenna had finally arrived.

loaded dipole program for the TRS-80

Dick's program was designed to be

used with an Apple II computer, but my good friend Dick Rasor, W6EDE, easily converted it for use with the TRS-80 (**fig. 1**). A little work with the program showed up some interesting aspects of the loaded dipole which previously had been obscured by the difficulty of the mathematics. These difficulties were now reduced to punching a few computer keys!

An illustration of the loaded dipole is given in fig. 2. For simplicity, the loading coils are located midway down the arms of the dipole: early experiments indicated this was the best place to put a loading coil if the assembly was to avoid becoming mechanically too complex.

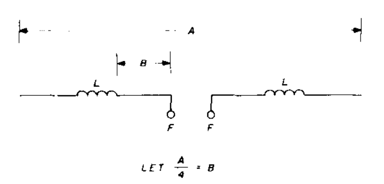


fig. 2. Loaded dipole with loading coils placed one-half the distance from the feedpoint (F-F) to the end. Distance A is one-quarter wavelength.

A computer run of the antenna design shows why coil placement is critical. **Fig. 3** plots coil placement against coil inductance. One limit on where the coil can be placed is seen at point 1, the feedpoint of the antenna. A feedpoint-loaded dipole places the coil at the point of maximum current, where the stored magnetic energy is high. A minimum value of inductance is required to establish resonance there, but — unfortunately — the portion of the antenna that does the most radiating is the portion with the maximum current. Winding it up into a coil reduces the radiation resistance, reduces bandwidth drastically, and leads to high antenna losses, principally because the coil will have relatively high loss no matter how well it is built.

Farther out along the antenna, the stored magnetic energy decreases and the inductance required in any coil placed there increases. At the same time, more of the high-current center section of the antenna is permitted to radiate. Antenna efficiency rises and the radiation resistance increases. Good!

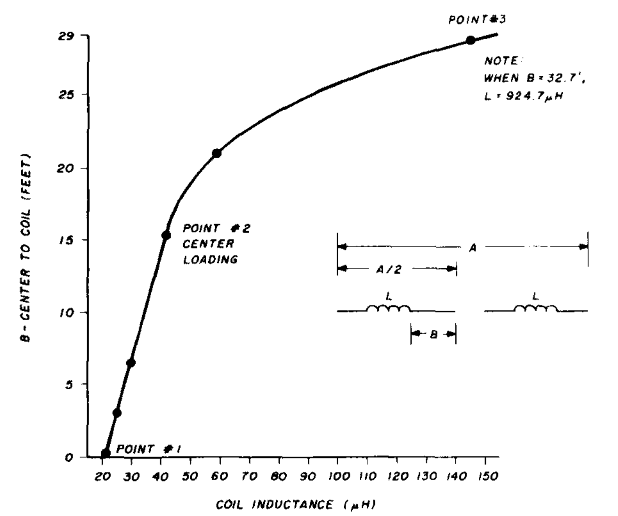


fig. 3. Inductance increases as the coil moves outward from the center of the loaded element. The increase in inductance is linear until coil reaches the center point of the element, and then it increases rapidly approaching the tip. If the coil is placed at the tip, inductance would theoretically have to be infinite. Coil loss increases with inductance, and point 2 on curve represents a practical compromise. Point 1 is for base loading.

But observe what happens when the coil passes the center point of the dipole leg (point 2). Now instead of increasing somewhat linearly with distance, the coil inductance increases rapidly. When the coil is placed near the end of the antenna (0.3) the required inductance value is more than seven times the value required for center (base) loading, and more than three times the value required when the coil is placed near the midpoint of the element.

It is tempting to place the loading coil near the tip of the antenna element; then, the whole element section has a high value of current in it, and this is thought best for antenna efficiency. But imagine a 925-μH inductor at 3.5 MHz. It would be four inches in diameter and have nearly two-hundred turns on it. The length would be over a foot, depending upon wire size. Placing such a coil at a high potential point in an antenna would result in fireworks: corona and brush discharge would occur with but a few watts of power applied. (And the coil would probably burn up after dust and dirt collected on it. In fact, all that would be required to do the job would be fog or rain.)

Fig. 4 shows the inductance of coils needed to make a half-size dipole for the various high-frequency bands. Although the antenna is not thereby reduced to its theoretically smallest size, this will show how an antenna can be cut fifty percent in size and still do a good job.

The computer printout that derived **fig. 4** was based on an antenna using No. 16 wire for the coils and flattop. If a larger size wire is used, the tip sections of the antenna should be shortened a few inches (this is not critical).

With this data, a short dipole for 3.8 MHz works out to be about 61 feet 6 inches (18.94 m) long. The

loading coils are each 40.1 μH, and they are placed 15 feet 4½ inches (4.69 m) from the center of the insulator.

How do you wind a 40.1-μH coil? There's a computer program for that, no doubt, but I don't have one at hand. However, the simple formula shown in **fig. 5** will do the job.

feeding the loaded dipole antenna

With a portion of the antenna wound up into a loading coil (L_1), the radiation resistance of the antenna drops drastically. For this design, the feedpoint resistance (composed of the radiation resistance plus the loss resistance of the coils) is about 22 ohms. This figure varies with height of the antenna above ground. Taking this value as par, the inductor-match system (hairpin match) developed by Gootch, Gardner, and Roberts will do the job.[5] For this antenna design, an inductor of about 44-ohms reactance (L_2) is placed across the antenna feedpoint. At 3.8 MHz, this corresponds to a coil of 1.86 μH. The reactance of the coil is derived from the graph in **fig. 6**.

Since the inducto-match is a simple L-network, the capacitive portion of the circuit is achieved by slightly shortening the antenna. Four inches off each end is about right, and the completed antenna is shown in **fig. 7**.

complete TRS-80 program for all bands

Using this information as a starting point, some smart computer programmer can develop a complete TRS-80 program which includes the design of the inducto-match coil. And, in addition, the program might be further expanded to include large-diameter elements. This will permit vertical antennas composed of aluminum tubing to be quickly designed for the lower frequency bands. I'll be happy to hear from anyone who completes this task.

no-code ham license?

A lot of flak is flying around about the so-called no-code license proposed by the FCC. The arguments against a no-code license seem to fall into two categories:

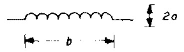
1. I had to pass a code test, so why shouldn't the next guy?

2. A no-code license will open the door to CB operators, who will ruin the ham bands.

I won't comment on the first argument, or the accompanying argument over tradition; others can fight that battle. But I would like to discuss the second argument that a no-code license would open the door to CB operators, who will ruin the ham bands.

frequency	total length feet	center to coil feet	coil inductance μH	wire diameter inches
1.82	128.57100	32.14290	91.88680	0.058
3.51	66.66670	16.66670	43.87170	0.058
3.80	61.57900	15.39470	40.10240	0.058
7.15	32.72730	8.18182	19.53040	0.058
10.11	23.14540	5.78635	13.12140	0.058
14.17	16.51380	4.12844	8.88137	0.058
18.11	12.92100	3.23026	6.67597	0.058
21.20	11.03770	2.75943	5.55306	0.058
24.95	9.37876	2.34469	4.58680	0.058
28.60	8.18182	2.04545	3.90515	0.058

fig. 4. Computer-derived table of the inductance values of coils needed to make a half-size dipole.



$X_L = 2\pi fL$

WHEN f = MHz AND L = μH, THEN

$$L = \frac{X_L}{6.28f}$$

$$L(\mu H) = \frac{a^2 n^2}{9a + 10b}$$

n = NUMBER OF TURNS

fig. 5. Formula for calculation of small close-wound coils for a given value of reactance, when f in MHz is known.

Perhaps this is true. But perhaps the CBers don't want to work in a VHF ham band! How about that!

It is very instructive to tune across the hf spectrum with an "all-wave" receiver. Anyone who does will note that there's a tremendous amount of illegal sideband activity between 26.2 MHz and 27.99 MHz. I believe there are more unlicensed stations in this portion of the spectrum than there are licensed stations in all the ham bands, at any one given time. This portion of the spectrum is jammed with thousands of signals.

These pirate operators are called "CBers." Perhaps this is an inaccurate epithet. I doubt if the majority of them have a CB license, and I prefer the term *pirate*. That does not imply CB operation. Be that as it may, the point I am bringing out is that these pirates operate wherever they wish, using modified ham gear. If they want to work on 144 MHz, or 220 MHz, they will do so — regardless of whether or not a no-code license exists.

When the sunspot count drops and the MUF falls, the 11-meter region will be barren of long-distance contacts. What will the tens of thousands of pirate operators do then? Go to the new no-code ham license? I doubt it.

Already many pirate operators in Europe are using the 6.6 to 6.8 MHz portion of the spectrum for SSB operation. The pirates tend to avoid the ham bands. They operate in the large

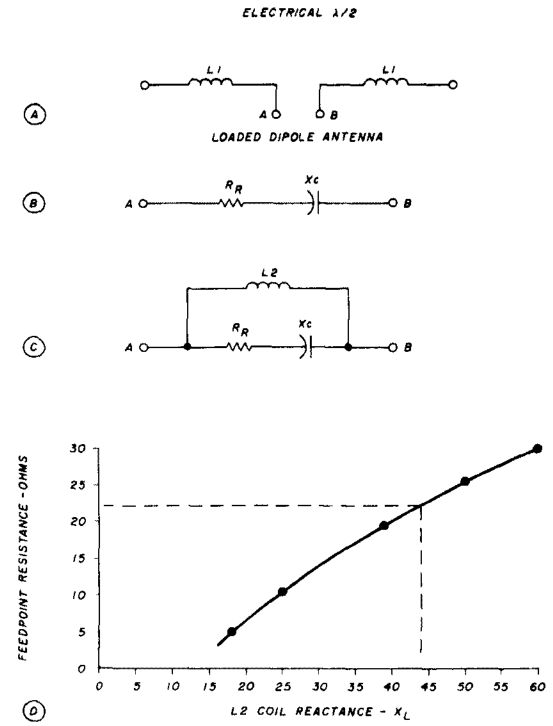


fig. 6. The coil-loaded dipole forms a portion of a network whose input impedance over a small frequency range is close to 50 ohms. The loaded dipole, *A*, has a low value of radiation resistance and loss resistance, which appears at feedpoint A-B. This low impedance can be made part of an equivalent parallel resonant circuit in which the total feedpoint resistance appears in series with the reactive branch of the circuit. *B*: The input impedance of such a circuit varies nearly inversely with the radiation resistance of the dipole, thus the low value of feedpoint impedance can be transformed to a larger value to match the line impedance. *C*: The dipole appears as a capacitive reactance by shortening the element past its resonant length. The inductor L2 consists of a small coil placed across the terminals of the dipole. The reactance of the matching coil is a function of the feedpoint resistance of the antenna. *D*: The dashed line is the example given in the text. Apply reactance value to formula given in *fig. 5*.

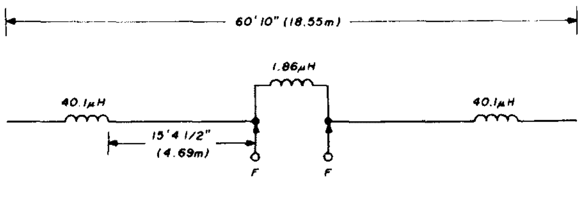


fig. 7. Compact, coil-loaded dipole for 3.80 MHz. Tip length is adjusted for minimum SWR at design frequency. Coil is wound with No. 16 (M1.3) wire per data in *fig. 5*.

spaces in the commercial and point-to-point regions, where few if any signals exist.

I say that the fear that pirates might invade ham radio via the no-code license is unfounded. They will come in only if they want to, regardless of the license structure, and my prediction is that there are more attractive places in the radio spectrum for them to occupy than a ham band. So I don't see the foundations of ham radio crumbling because a no-code license is introduced.

As time goes on the number of pirate stations will *increase*, because the various communications authorities throughout the world seem powerless to stop them. A few pirates will inevitably invade the ham bands from time to time, but this will have nothing to do with the Amateur Radio licensing structure. The problem has been swept under the rug up to now, yet it increasingly involves all the radio services. Pirate radio includes illegal broadcasting on medium and shortwave and VHF. In Europe, pirate broadcasting clogs the fm band and the quieter broadcast band channels. There are pirate television stations in operation in Europe, and Central America is full of illegal broadcasting. So far, radio hams are lucky; little of this trash has fallen in their bands. The pirates prefer to go where they can operate under less scrutiny than in a busy ham band.

So don't worry about a VHF no-code license. The pirate operators have more alluring possibilities open to them than competing with hams in a short-range, line-of-sight service.

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ham radio

a microprocessor repeater controller

A versatile controller for two repeaters

Our radio club recently relocated its 2-meter repeater to a site with a much higher antenna. The repeater committee decided to make major improvements in the control system to accommodate this move. We had a 220-MHz repeater also under construction, and would need a controller for that system as well.

The original controller was a small, wire-wrapped board using 556 timers, some counters, and a ROM for the CW identification. Remote control was by phone line and was not sophisticated. Past experience with this system indicated that adding any simple function would be a major task. Microprocessor enthusiasts in the club had the solution: build *one* microprocessor-based controller for both bands!

The final design may be expanded upon easily. In addition to the hardware description, I would like to

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share some of our thoughts and decisions that went into creating the final design.

the design approach

Deciding to use a microprocessor as a controller was easy. In the long run it would be cheaper, and it's easier to add features by reprogramming than to add separate pieces of specialized hardware. Some new circuitry would be needed as features were added, but such circuitry would be simple interfaces.

Reliability would be good, thanks to the high reliability of digital circuits and the lower parts count per function (compared with standard small and medium-scale integrated circuits). Two decisions had to be made: which microprocessor to use, and what features to include in the new controller.

selecting a microprocessor

The microprocessor we finally decided on would have to be easy to program in assembly language, have a simple input/output (I/O) structure, and be supported by good development software. The microprocessor instruction set should be able to handle reentrant programming, allowing one program module to share multiple data sets.

The Intel 8080, Zilog Z80, Motorola 6800, and Texas Instruments 9900 microprocessors were all candidates for our application. The 8080 or Z80 at first appeared to be the best choice. A friend had built an 8080 controller six years ago for the WR8ANW repeater in Columbus, Ohio. The program listing used for that controller was available and could have been converted for our needs. Several club members had 8080 systems that they used for software and hardware development. A major drawback of the 8080 was its I/O structure and the difficulty of writing clean, reentrant code for it.

The Z80 has few of the shortcomings of the 8080. It can set and test single bits in operands, has an indexed addressing mode, and allows I/O port addresses to reside in one of its internal registers. Reentrant programming is easier with it. Unfortunately, none of the club members had Z80 support software at that time.

The 6800 was not really in the running. None of the club members were familiar with it; we would be starting from scratch. This doesn't mean the 6800 won't work for this application. The WR8ANW repeater group, mentioned earlier, has completed a 6800-based controller.

The TMS9980 was our final choice. It is easy to write reentrant code for the 9900 family since any register may be an index. Interrupts are easily handled. Since all general purpose registers are in memory, the only registers saved on interrupt are the program counter, status register, and workspace pointer. These three restore automatically after interrupt servicing, reducing the programming load. And, support software which became available to the club on a larger 990 minicomputer proved to be a valuable tool when it came time to assemble and edit the controller programs.

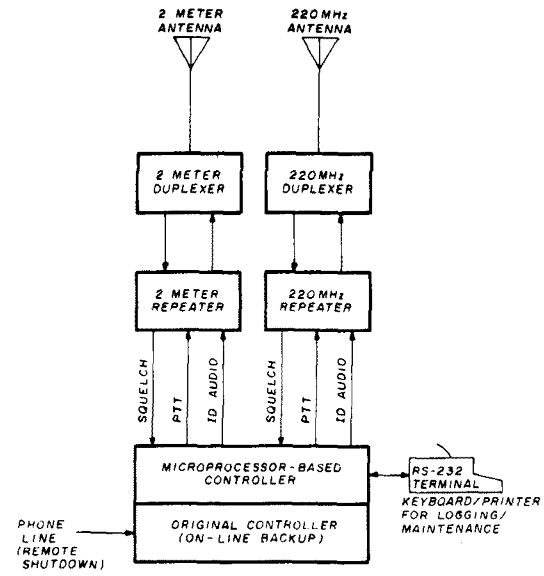

fig. 1. Overall system block diagram of the K5OJI double-frequency VHF repeater. New controller includes a keyboard/printer terminal for logging and maintenance. Original controller on-line for backup. Telephone remote shutdown disables entire system.

table 1. CRU address decodes in TMS9901 PSI. Addresses are in hexadecimal.

CRU hardware address	R12 contents	device
00	00	–
20	40	–
40	80	9902(A)
60	C0	9902(B)
80	100	9901
A0	140	–
C0	180	–
E0	1C0	–

a choice of features

A list of the minimum functions required for our application was drawn up. These included CW identification, a variable time-out timer, a beep to indicate time-out, timer reset, and a status-logging rou-

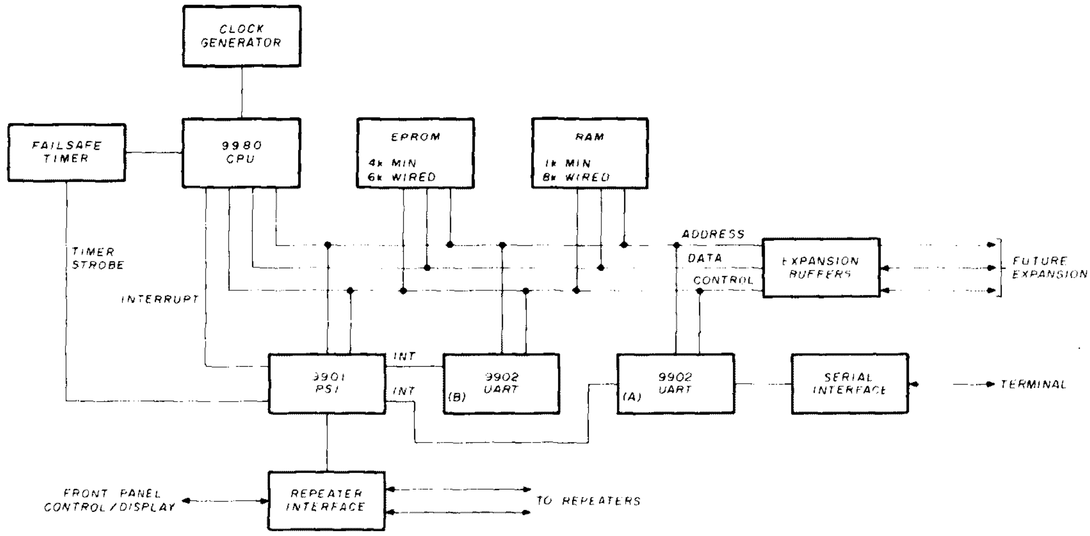

fig. 2. Block diagram of new microprocessor controller. Architecture allows expansion for future functions.

tine to print hourly status reports on a terminal. The time of day was added to the CW ID since there would be a counter keeping track of time in the program.

Keeping the original controller, modified to operate as a backup, would retain telephone line shutdown with the ability to disable the repeaters regardless of which controller was operational.

Fig. 1 is the repeater system block diagram. It was constructed so that adding new features would cause only a few hours of downtime. New programming may be installed while the backup controller handles the repeater. Some of the new features include a tone decoder, a modem for RTTY I/O and control, and even a voice synthesis module.

We had defined the general system; features were chosen and the microprocessor would use TMS9900 family components. This left only the hardware details to design.

build or buy?

Texas Instruments makes several single-board microcomputers. The TM990/100 and TM990/180 boards have a small prototyping area where additional interface circuitry can be built. Each has plenty of onboard EPROM and RAM for program operation.

The final program would be burned in the EPROM, but I wanted to put the program in RAM first to do the final debugging and possible *patching*. The temporary RAM test-space would be free after program verification. The free RAM could then be used for other functions, perhaps as a message storage area for RTTY users. The only way to get enough checkout RAM with the TM990 boards is to add at least one additional board.

Designing and wire-wrapping a single board with enough memory and I/O components to meet the basic criteria seemed the best way to proceed. It would include enough circuitry to bring address, data, and control lines off the board for later additions. Later features could be added using separate boards.

Fig. 2 is the single-board controller block diagram. Memory and I/O addressing are similar enough to the TM990 board series to allow using the TIBUG™* monitor ROM for program check-out, and also allow the final debugged control program ROMs to be installed and run on the TM990 board.

solidifying the design

Figs. 3, 4, and **5** are the schematics for the controller board. Signal mnemonics connect the three main schematic groups. Two edge connectors, P1 and P2, connect the controller to the rest of the system. Details begin with **fig. 3**. The controller chassis is seen in **photo 1**.

The Central Processor Unit (CPU) is a TMS9980 with an 8-bit data bus and addressing to 16K (16,384) bytes, more than sufficient for this application. CPU clock frequency is 10 MHz, from the crystal-controlled inverter oscillator in U8. External device clocking is available at U20-22, marked $\overline{\text{Ø3}}$.

The CPU resets by interacting with peripheral interfaces, shown in **fig. 5**. Power-on reset for these in-

*TIBUG is the Texas Instruments debugging utility.

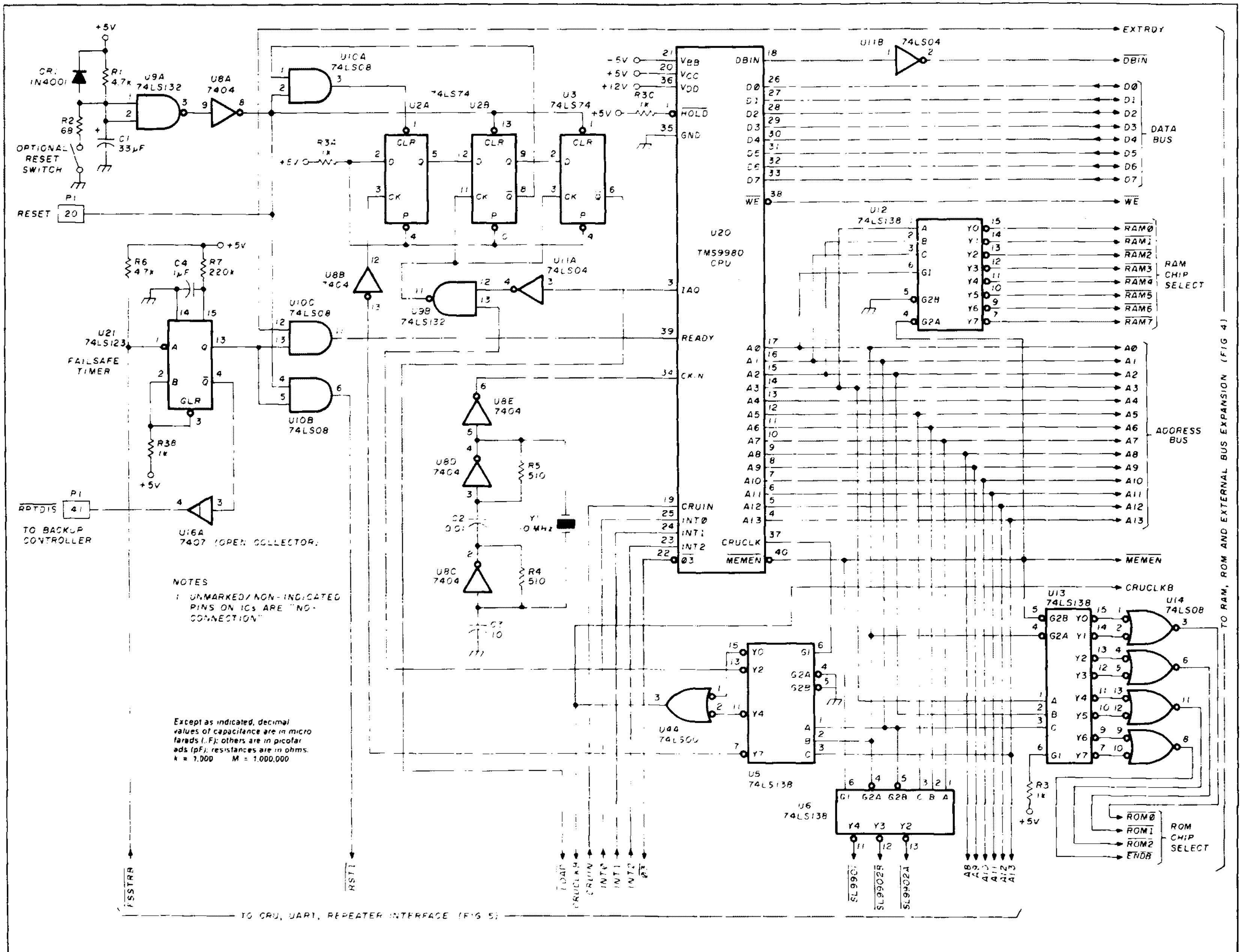

fig. 3. Schematic of central processor, address decoder, clock and reset circuitry. P1-4 face in fig. 5.

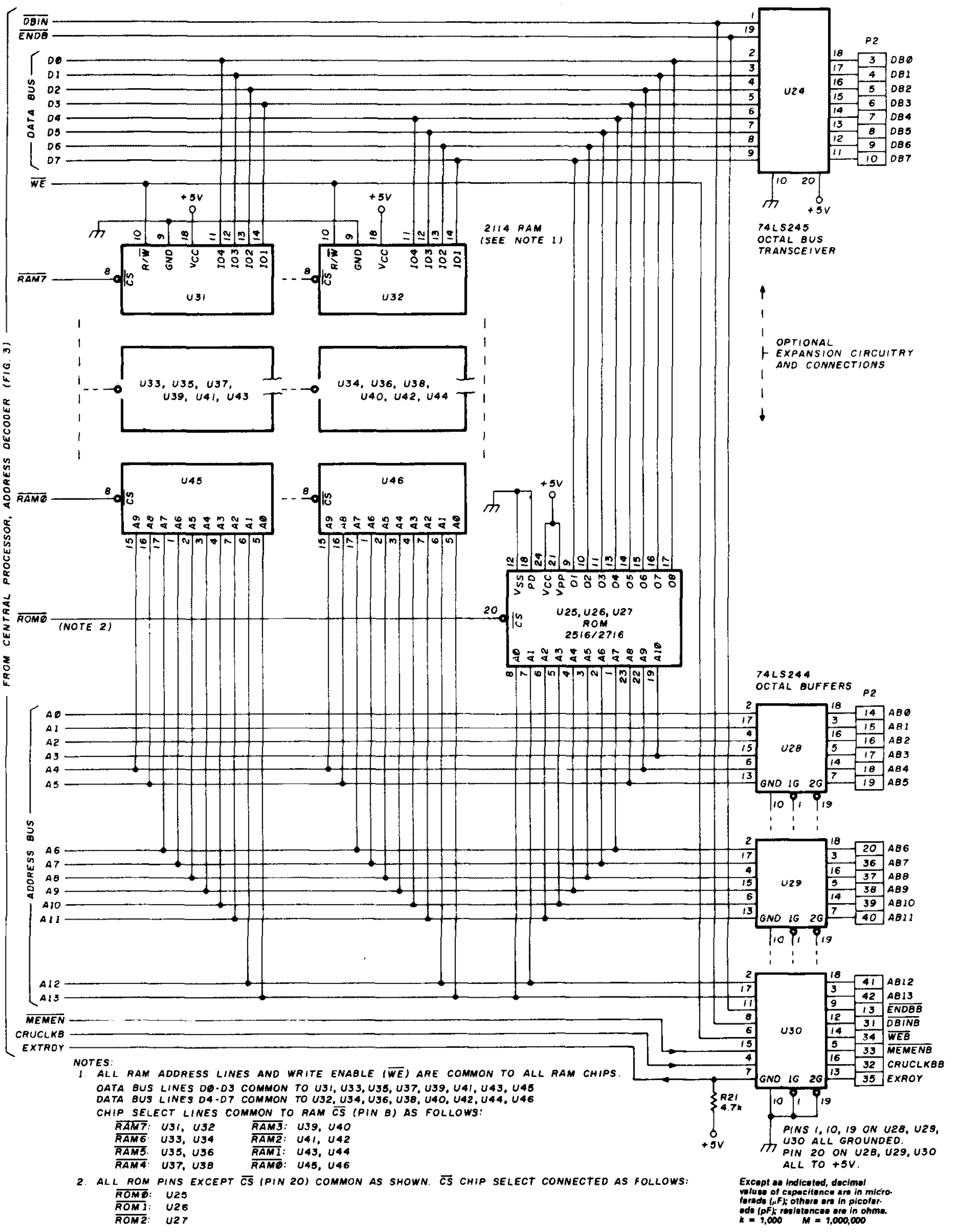

fig. 4. Memory and expansion bus interface. Bus buffers and transceiver may be added later. U31 and U32 are minimum requirements for RAM, U25 and U26 minimum for ROM.

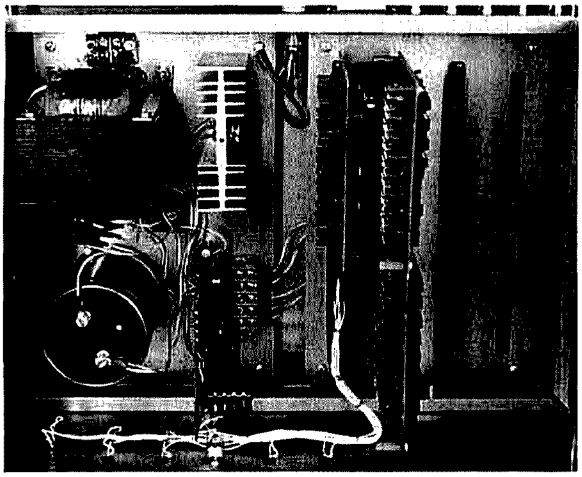

photo 1. Top view of controller chassis with power supply. New controller is long board in second slot. Stand-offs protect wire-wrap pins on IC sockets. First slot contains old controller, smaller board toward chassis rear. Small board in first slot toward front panel contains LED resistors and wire connections to front panel. Empty slots are for future expansion. Edge-connector socket pairs wired in parallel with P1/S1 toward front.

terfaces is provided by Schmitt-input gate U9A, R1, C1 through U8A and U10B to the $\overline{\text{RSTI}}$ line. A normally open reset switch may be added for testing. $\overline{\text{RESET}}$ at P1-20 to reset future external circuits.

The failsafe timer one-shot at U21 is re-triggered by the program through $\overline{\text{FSSTRB}}$ every 16.7 milliseconds. As long as the controller is operational, $\overline{\text{RPTDIS}}$ at P1-41 remains low and disables the backup controller. Controller failure will make $\overline{\text{RPTDIS}}$ high and enable the backup; a TTL pull-up resistor is located in the backup controller.

Flip-flops U2 and U3 generate proper $\overline{\text{LOAD}}$ timing for the interface chips with the help of decoder U5 and gate U9B. The ready signal input to the CPU (U20-39) must be high for normal operation with memory; the low state causes a CPU wait mode, to allow for access with slow memory. AND gate U10C keeps the ready signal high via the failsafe one-shot and expansion signal EXTRDY.

RAM is organized in 1K banks, chip-pairs selected by U12. ROM is in 2K banks, selected by U13 and U14. $\overline{\text{ENDB}}$ is a fourth 2K bank select for expansion. U6 selects the interface chips and is wired for selecting one of three 32-bit CRU I/O bit groups. Addressing is detailed in the last section.

memory and bus expansion

Fig. 4 is a simplified memory schematic. Static RAM uses 2114 devices having a 1K by 4-bit structure. Address lines A4 to A13 and write enable WE are common to all RAM chips, but data bus lines must be split as indicated. RAM chip select lines $\overline{\text{RAM0}}$ to $\overline{\text{RAM7}}$ must be common only to a pair of 2114s.

All ROM pins except chip selects $\overline{\text{ROM0}}$ to $\overline{\text{ROM2}}$ are common. Either a 2516 or 2716 $\overline{\text{EPROM}}$ may be used for ROM, but there is a slight programming difference between the two. Both RAM and ROM may use 450 nanosecond access time devices.

A minimum system must have U25, U26, U31, and U32 installed. All memory sockets are wired for ease of check-out. The board in photo 2 shows 4K RAM installed for program verification. The memory map is seen in fig. 4A.

Bus transceiver U24 and bus buffers U28 to U30 are needed only if expansion is considered. R21 must remain to hold EXTRDY high if U30 is removed.

talking to the rest of the system

The TMS9901 Programmable Systems Interface (PSI) is the key device in fig. 5. It provides interrupt masking, priority encoding, I/O ports, and an interval timer in one package. It also handles interrupts from the TMS9902 Universal Asynchronous Receiver/Transmitter (UART) at U17 and U18.

The 9901 communicates with the CPU through the CPU's communications register unit (CRU), an internal serial interface within the 9980. (The CRU operation is covered briefly later in this article, but the reader is referred to the reference for detail.)

The open-collector buffers to the repeaters and

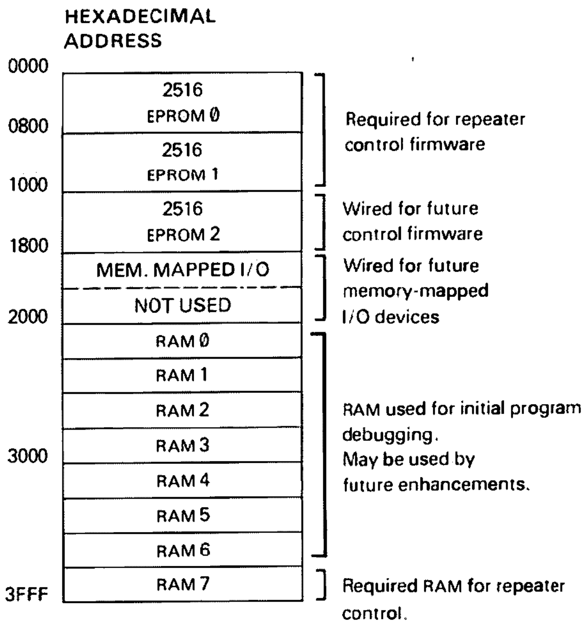

fig. 4A. Memory map of controller. Monitor ROM located in $0000 to $07FF address. Minimum RAM in address location $3D00 to $3FFF ($ = hexidecimal).

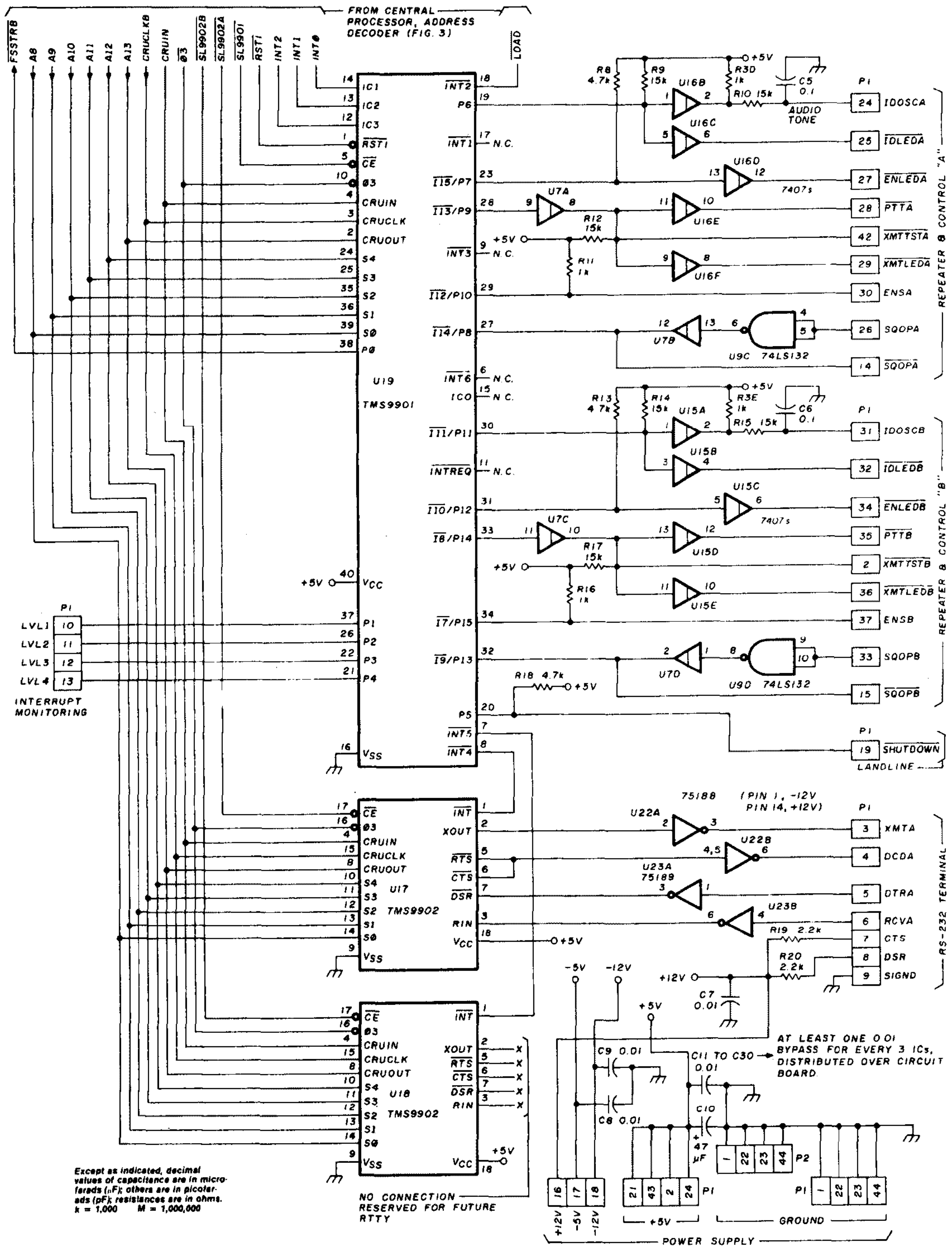

fig. 5. Interface circuitry and power supply connections. UART in U18 used mainly for internal timer. RC filter for CW ID tone (IDOSC) should be isolated to minimize digital noise. A minimum of one 0.01 μF bypass per three ICs is recommended on entire board.

front panel controls are identical circuit groups to each repeater. Mnemonics for the signals have an A suffix for the 2-meter repeater, a B suffix for the 220-MHz repeater. Direct repeater controls are $\overline{\text{PTT}}$ (push to talk), IDOSC (ID tone or 'oscillator'), and SQOP (squelch open). Other signal lines in each group refer to the front panel controls and indicators shown in **fig. 6A** and in **photo 3**.

$\overline{\text{PTT}}$ is low to transmit. Pull-up for the open-collector buffers (U16E and U15D) is provided within the repeater chassis. The CW identification tone is provided by programming the first-level interrupt period of 512 microseconds for a square-wave frequency just under 1 kHz. RC filtering at the IDOSC output produces a triangular waveform with an amplitude of about 5 volts peak-peak.

Remote shutdown is common to both controllers, but the direct telephone interface is within the old controller. Two rings on the landline will cause $\overline{\text{SHUTDOWN}}$ to go low, disabling the main controller. $\overline{\text{SHUTDOWN}}$ is TTL-compatible but requires R18 to hold U19-20 high when the backup controller is removed.

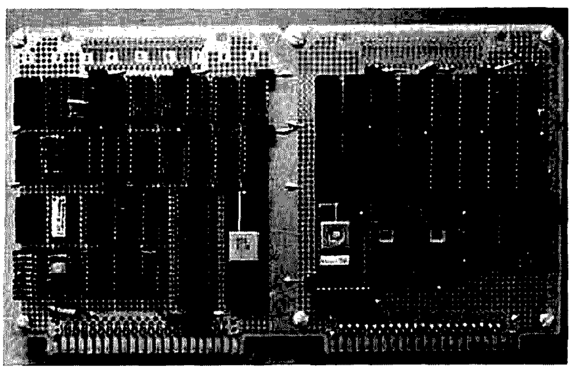

photo 2. Photo of controller board made just before final circuit freeze. All ICs are socketed on prototype board. Except for supply bypass capacitors, all discrete components mounted on DIP plugs. Number labels were construction references.

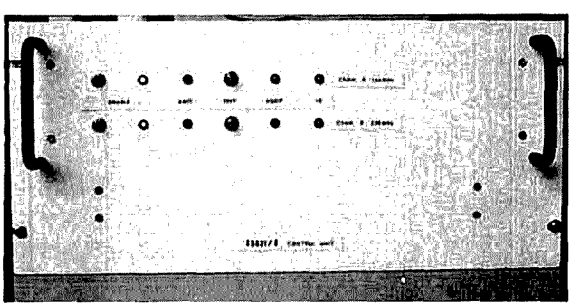

photo 3. Front view of front panel controls and indicators.

The RS-232 terminal connections (completed in **fig. 6C**) use high-voltage buffers in U22 and U23 for an ASR-733 terminal. Other devices can be used to interface the UART at U18. The terminal is connected directly to the new controller, and not used in the backup.

The power demand of the single-board controller is 3 A at +5 Vdc, 2 mA at −5 Vdc, 0.2 A at +12 Vdc, and 0.1 A at −12 Vdc. The +5 Vdc supply demand is dependent on the amount and type of memory. A well-regulated supply should be used, but the current should be calculated for your own configuration.

manual control and indication

The front panel controls are not an absolute requirement, but do provide local control for testing and a quick indication of operation.* The 2-meter control and indication is shown in **fig. 6A**; the 220 MHz arrangement is identical except for interconnecting pins.

The condition of the ENABLE switch is periodically read by the program. Switch status, shutdown signal, and a flag in memory will determine if the particular transmitter should be turned on when requested. The ENABLE status is displayed by the program as a check of all conditions.

The XMIT display lights up whenever a repeater is transmitting. The TEST switch controls two methods of transmit: manual — without microprocessor control — if the switch is held to the left, or simulation of squelch-open with processor control if it is held to the right. The SQOP display indicates the latter simulation, or normal squelch-open condition, of the repeater.

The ID LED is driven from the same source as the audio tone. Since this signal is a fifty-percent duty cycle, the current limiting resistor is smaller, creating a more uniform brightness.

Four keyboard commands are recognized. An operator can type U on the terminal to update the time, T to print current program time, M to modify the clock, and S to print the current system status. Other entries are ignored. The time correction is the number of seconds to be added to the internal clock each day; there is no provision for tweaking the system clock frequency.

construction

The controller was wire-wrapped on a prototype board, as shown in **fig. 7** and **photo 2**. Bypass capacitors for the +5 V supply line were soldered directly on the board, one for every three ICs and one

*User-friendly controls and terminal commands benefit the non-computerist in your repeater committee.

for every two memory chips. All other discrete components mount on DIP plugs.

A 12 × 17 × 2 inch (30 × 43 × 5 cm) Bud chassis is bracket-mounted to the rack panel. Two 7 × 11 inch (18 × 28 cm) aluminum plates hold the power supply and four pairs of card edge connectors. All interface connectors, the line fuse, and switch are mounted on the rear face of the chassis.

programming and checkout

The program was coded in short routines, most containing less than fifty lines. The code is heavily commented to facilitate debugging and to provide good documentation. Documentation is essential if you want anyone, even the programmer, to understand the program at a later date.

The program was initially programmed into the EPROM and installed on the board. A short routine was executed to move the program from ROM into RAM. Execution from RAM was under control of the monitor, allowing correction and patching. The monitor used was TIBUG™.

The EPROMs are re-programmed after checkout so the program can execute from a ROM address area rather than RAM.

Hardware and system checkout procedure used the front panel TEST switch to simulate the receiver squelch-open signal until most of the program bugs were found. Later, the $\overline{\text{PTTA}}$ line was jumpered to SQOPB and $\overline{\text{PTTB}}$ was jumpered to SQOPA; with both channels enabled, the controller would alternately transmit on 2 meters and 220 MHz. We ran the controller only in this mode for several days in the presence of the club's HF and repeater equipment to verify that the controller was rf compatible. No interference was observed. A typical printout is seen in **fig. 8**.

history

Total construction time for this project was approximately four months. Most of the board wiring and program design was completed during a two-week vacation. The most time-consuming task was packaging the controller.

The controller was installed in the K5OJI repeater in January, 1981. Up to the time of this writing,

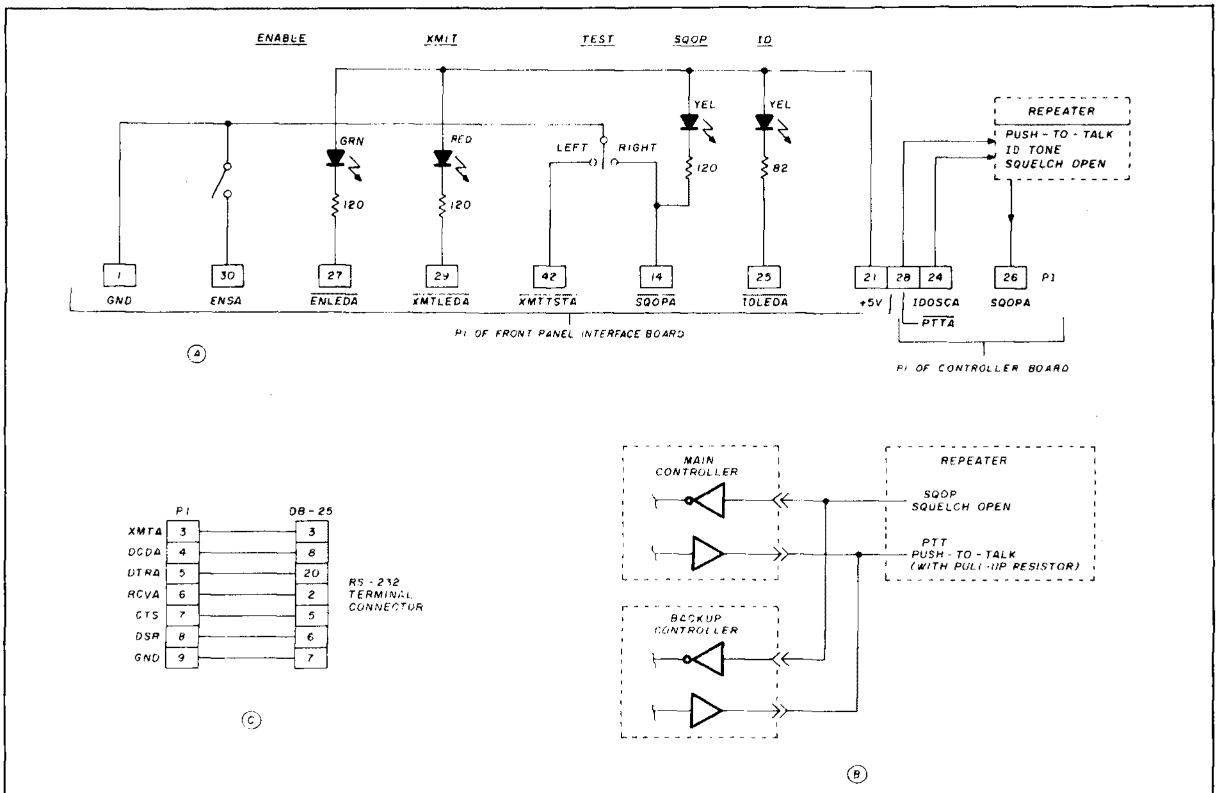

fig. 6. Front panel controls and indicators for 2-meter repeater in A. 220 MHz front panel and repeater connections are identical except for pin numbers. Underlined names are identical to front panel photo. B indicates parallel connection scheme for backup controller. C gives RS-232 serial interface pin connections to main controller.

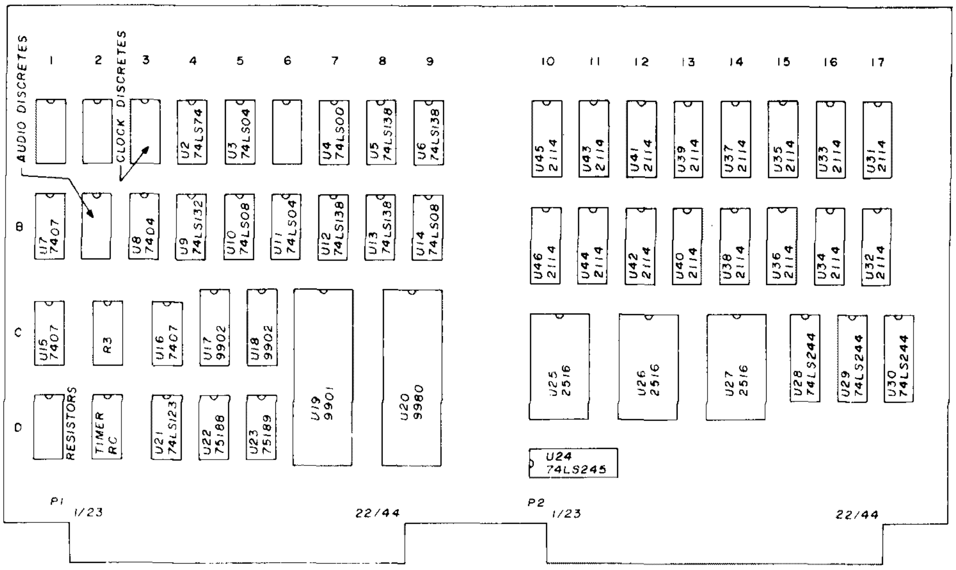

fig. 7. Location of ICs on controller board. Edge connections are double side, and pins etched in board. Letters and numbers used for construction reference. Two RAM chips in U31, U32 is the minimum requirement but photo indicates 4K RAM population up to U38 for development purposes.

about two years, we had only one failure due to bad memory chips. Since the backup controller picked up when the main unit failed, the repeater was never off the air. The bad chips were quickly located and replaced.

The controller is reliable, expandable, and relatively simple. It can be made on a prototype board or it may be an adaptation of commercially available microprocessor boards. Hardware and software is designed so other features may be added easily.

Based on observation of microprocessor loading, the controller should be able to control three repeaters simultaneously. The Level 1 interrupt is the heaviest CPU load and provides the ID tone; a separate hardware oscillator will relieve much of the first-level interrupt handling.

This project would not have been possible without the help of WB8CEB for most of the program editing and N5JS and AJ5L who maintained the rf portions of the repeaters.

A listing of the control program is available on an 8-inch CPM™ compatible disk available from the author for $15.00. This disk contains the program listing and an object file for programming EPROMs. The disk is single-sided, single-density and the program uses 26 sectors at 128 bytes per sector.

```
ENTER CURRENT TIME HHMM: 2040
K5OJI RPT TIME IS 2041
TIME=2100 XMT=1409 OID=06 TMO=01 IDC=13 KCT=07
TIME=2200 XMT=0112 OID=05 TMO=00 IDC=07 KCT=00
TIME=2300 XMT=0044 OID=01 TMO=00 IDC=03 KCT=01
TIME=0000 XMT=0000 OID=00 TMO=00 IDC=00 KCT=01
TIME=0100 XMT=0031 OID=01 TMO=00 IDC=02 KCT=00
TIME=0200 XMT=0000 OID=00 TMO=00 IDC=00 KCT=00
TIME=0300 XMT=0000 OID=00 TMO=00 IDC=00 KCT=00
TIME=0400 XMT=0000 OID=00 TMO=00 IDC=00 KCT=00
TIME=0500 XMT=0000 OID=00 TMO=00 IDC=00 KCT=00
TIME=0600 XMT=0000 OID=00 TMO=00 IDC=00 KCT=00
TIME=0700 XMT=1660 OID=04 TMO=00 IDC=09 KCT=03
TIME=0800 XMT=0277 OID=09 TMO=00 IDC=14 KCT=03
TIME=0900 XMT=0837 OID=05 TMO=00 IDC=09 KCT=00
TIME=1000 XMT=0000 OID=00 TMO=00 IDC=00 KCT=00
TIME=1100 XMT=0000 OID=00 TMO=00 IDC=00 KCT=00
```

fig. 8. Typical printout of part of one day's operation.

for the computer technician

Computer technology is a specialized area. Some explanations and technical arguments follow which will serve the needs of the computer specialist who undertakes this project. ham radio *cannot take sides in programming techniques, but a strong relationship between hardware and software is integral to the successful design of this system, and the computer technician should be aware of this before beginning the project. Editor.*

Reentrant programming is sometimes confused with recursive programming; we offer the following abbreviated definition from Granino Korn's *Microcomputers for Scientists and Engineers*:

"A special case occurs where a subroutine is inter-

table 2. TMS9901 PSI bit assignments.

select bit	CRU read data	CRU write data
0	Control bit (1)	Control bit
1	INT1-/CLK1 (2)	MASK1/CLK1 (3)
2	INT2-/CLK2	MASK2/CLK2
3	INT3-/CLK3	MASK3/CLK3
4	INT4-/CLK4	MASK4/CLK4
5	INT5-/CLK5	MASK5/CLK5
6	INT6-/CLK6	MASK6/CLK6
7	INT7-/CLK7	MASK7/CLK7
8	INT8-/CLK8	MASK8/CLK8
9	INT9-/CLK9	MASK9/CLK9
10	INT10-/CLK10	MASK10/CLK10
11	INT11-/CLK11	MASK11/CLK11
12	INT12-/CLK12	MASK12/CLK12
13	INT13-/CLK13	MASK13/CLK13
14	INT14-/CLK14	MASK14/CLK14
15	INT15-/INTREQ	MASK15/RST2- (4)
16	—	Fail-safe strobe
17	—	LEVEL 1 INTERRUPT INDICATOR
18	—	LEVEL 2 INTERRUPT INDICATOR
19	—	LEVEL 3 INTERRUPT INDICATOR
20	—	LEVEL 4 INTERRUPT INDICATOR
21	Remote shutdown	—
22	ID Ch A	ID Ch A
23	—	ENABLE Ind Ch A
24	SQ OP Ch A	—
25	—	XMIT Ch A
26	ENABLE SW Ch B	—
27	ID Ch B	ID Ch B
28	—	ENABLE Ind Ch B
29	SQ OP Ch B	—
30	—	XMIT Ch B
31	ENABLE SW Ch B	—

(1) 0 = interrupt mode, 1 = clock mode
(2) Data present on INT pin (or clock value) will be read regardless of mask value.
(3) In interrupt mode writing a 1 into mask will enable interrupt, a 0 will disable the interrupt.
(4) Writing a zero to bit 15 while in the clock mode executes a software reset of the I/O pins.

rupted and the interrupt calls the *same* subroutine. A program may fail on return from interrupt. Subroutines designed to work properly on interrupt and restoration from interrupt are called 'reentrant.' A good way to obtain reentrant subroutines is to provide temporary storage of addresses and register contents in 'stack' storage. Real-time computation with many interrupt-driven segments make reentrant programming desirable."

When many repeaters need be controlled, the only additional software necessary should be new parameter tables and calls to the routines handling data in these tables. Not only should the data manipulation instructions be reentrant, but so should I/O instructions; controlled devices will not always have the same I/O addresses.

The I/O structure of the 8080 does not lend itself to reentrant programming. I/O routines must be programmed once for each channel, and you must decide which piece of code to execute, or the code must be written to be self-modifying: the program modifies the instruction set about to be executed before entering the set. The instruction must reside in RAM to be self-modifying. The I/O of the 8080 must transfer eight bits at once, which requires extra logical instructions. This means that the bits which control the repeater must be set, reset, and tested, or only one function can be assigned to each I/O port.

While you need subroutines to load and test an 8080 memory location, a single 9900 instruction performs the same function. The 9900 I/O structure lends itself to reentrant programming. The 9900, through its CRU, may transfer from one to sixteen bits with a single instruction. This makes it suitable for multiple-control applications.

The address bus I/O address is generated by add-

ing the CRU bit address in the instruction to the contents of the CRU base register, one of the user-accessible registers. By setting base register contents differently for each channel, the same I/O instructions can be used to control the same function on different channels.

Since all general purpose registers are in memory, only the CPU program counter, status register, and workspace pointer need be saved during an interrupt. These are saved and restored automatically. The programmer does not have to keep track of which registers to save or restore.

The TMS9980 CPU is part of the 9900 family and uses the same instruction set. This class of processor differs from earlier designs and readers should refer to the reference material for exact details. The following will help you understand the CRU and how it is used in the K5OJI repeater.

understanding the CRU

The communications-register-unit uses a dedicated bit-addressable interface for I/O between the CPU and 9901, 9902 devices. The CRU interface in the system is the address bus and three signal lines: CRUCLK, CRUIN, and CRUOUT (multiplexed with address line A13 on the 9980). The 9901 and 9902s are enabled via U6 by address lines A0, A1, A5, A6, and A7 while address lines A8 through A12 select the single bit to be input or output. The CRU transfers data one bit at a time, serially, on the CRUIN and CRUOUT lines.

For output, the address lines are set to point to the desired output bit and that bit of data is put on the CRUOUT (A13) line. CRUCLK then clocks the data into the selected device. For input, the address lines are set to point at the desired input, then clocked into the CRU through the CRUIN line. There is no external signal to indicate when an input is read.

Table 1 lists the hardware and software addresses for the CRU. The 9901 occupies thirty-two bits of CRU input/output space and assignments are given in **table 2**. **Table 3** is a complete parts list for the controller.

Table 2 needs further explanation: bit 0 controls the mode of bits 1 to 15. If bit 0 is logic 0, the 9901 is in interrupt mode. Writing to bits 1 to 15 sets an internal mask for passing or ignoring an interrupt level.

The 9901 is in clock mode (internal interval timer) if bit 0 is logic 1. Writing to bits 1 to 14 loads a value into the timer's count decrementer. As the timer counts down to zero, an interrupt is issued and the timer resets to decrement value. Reading bits 1 to 14 will read the current value of the decrementer. Reading bit 15 inputs the status of the interrupt request while writing to bit 15 initiates a reset of input/output pins.

table 3. Controller parts list.

quantity	type	
1	TMS9980	U20
1	TMS9901	U19
2	TMS9902	U17, U18
2 min., 16 max.	2114	U31 to U46
2 min., 3 max.	2516	U25 to U27
1	74LS00	U4
1	74LS04	U11
1	7404	U8
3	7407	U7, U15, U16
2	74LS08	U10, U14
2	74LS74	U2, U3
1	74LS123	U21
1	74LS132	U9
4	74LS138	U5, U6, U12, U13
3	74LS244	U28, U29, U30
1	74LS245	U24
1	75188	U22
1	75189	U23
	resistors (all resistors are ¼ W, 10% unless otherwise specified)	
1	220 K	R7
1	68 ohm	R2
2	82 ohm	front panel
6	120 ohm	front panel
2	510 ohm (5%)	R4, R5
2	2.2 K	R19, R20
6	4.7 K	R1, R6, R8, R13, R18, R21
6	15 K	R9, R10, R12, R14, R15, R17
2	1 K	R11, R16
1 DIP array	1 K	R3 (Beckman 899-1-1.0 K)
1	220 K	R7
	capacitors (all capacitors are disk, 25 V min unless other specified)	
24	0.01 μF	C2, C7, C8, C9, C11 to C30
2	0.1 μF	C5, C6
1	1.0 μF	C4
1	10 pF mica	C3
1	33 μF	C1 (electrolytic, 10 V min.)
1	47 μF	C10 (electrolytic, 10 V min.)
1	CY-18 crystal, 10 MHz	Y1
1	1N4001	CR1
2	LED, green	front panel
2	LED, red	front panel
4	LED, yellow	front panel
2	switch, SPDT, momentary-off-momentary	front panel
2	switch, SPST	front panel

Bits 16 to 31 are for I/O, the majority directly interfacing with the repeaters. Writing a 0 and then a 1 to bit 16 will re-trigger failsafe one-shot U21. Re-triggering must occur at a 60 Hz rate.

Bits 17 to 20 are monitor output which indicates the level of interrupt processing. Entering an interrupt routine sets the appropriate bit for that interrupt level. Completing an interrupt resets the bit. Oscilloscope monitoring verifies the interrupt and indicates CPU loading for each interrupt time. The first three

interrupt levels are used here with the fourth level reserved for future use.

Bit 21 is an input for remote shutdown via telephone line through the old controller. The old controller will shut down through its own interface circuitry and a low state of SHUTDOWN will disable the new controller.

Bits 22 to 26 are I/O control for the 2-meter repeater ("A" suffix mnemonics) while bits 27 to 31 are identical in function for the 220 MHz repeater ("B" suffix).

Interrupt level 3 is internal to the 9901. Interrupt levels 4 and 5 are hardwired to the interrupt outputs on both 9902s. The 9901 will prioritize interrupts, outputting an interrupt code of 0 for highest priority and 15 for lowest priority. The 9980 CPU interprets levels 3, 4, and 5 as interrupt levels 1, 2, and 3, respectively.

Each 9902 UART is assigned thirty-two bits of CRU and each may cause an interrupt from four separate events. Repeater control uses only the interval timer interrupt. The second 9902 (U18) is used solely for the timer, but could be used for a second serial interface.

software

Author Warner claims that packaging the controller was the most time-consuming task and that software design was second. Judging from the 51 pages of program listing available, we might reverse that statement. The final excerpt contains some details on the program package.

The software design was to include as many features as possible and to break the program into small, easy-to-follow modules. These modules can be called by the appropriate interrupt processor module, depending on the desired frequency of execution. It would not be difficult to add modules for new features.

Modules communicate with each other (on the same and different interrupt levels) via semaphores, flags set in specific memory locations. Seven extended-operation (XOP) instructions are included for I/O with a keyboard/printer. The hardware will support a total of 16 XOPs, so users may add their own XOP routines.

Hardware reset causes an entry into the initialization section of the program. This initializes certain memory locations, I/O interfaces (including all interval timers), and the interrupt mask register in the 9901. Once accomplished, the program enables interrupts and begins execution of the program's polling loop.

The following program names are those included in the program. The interrupt level routines handle all the repeater control functions. Three levels of interrupts are used. Level 1 is highest and occurs when the 9901 interval timer decrements to zero. Program segment C04 generates the CW ID tone on a Level 1 interrupt. This will generate a 1 kHz tone for each repeater.

Interrupt handlers are similar. First the appropriate CRU output bit is set to indicate initiation of processing at the particular interrupt. Register 1, used as an index register, is loaded with the address of the parameter table for one repeater. The proper routines for that repeater are then called to operate on the parameters. When processing for one repeater is complete, Register 1 is re-loaded with the address of the parameter table for the other repeater, and the same routines are called again. When all processing for the interrupt level is complete, interrupt hardware is enabled for the next interval timer decrement-through-zero. The CRU bit, indicating process in operation, is reset and control returns to the interrupted routine.

Level 2 interrupt is caused by the interval timer in the 9902 at U17. This timer is set to decrement through zero every 4.7 milliseconds. The routine labelled C01 is executed on a Level 2 interrupt and forms the ID tone length and beep.

Main repeater timing occurs at Level 3, generated every 16.7 milliseconds. Some system functions, such as time of day and checking for remote shutdown, are executed only once per interrupt. All other repeater routines must be executed once for each channel. Routines R00, R07, and R09 are called only once while repeater routines R01 through R05, R08 are called twice.

When no interrupts are being serviced, the polling loop at I03 is operating. This loop checks for keyboard inputs and checks flags that indicate printout of an hourly repeater status. Once each hour the interrupt level routines move the hourly status for each repeater to a print buffer, clear the next hour's status, and set a print request flag. The polling loop checks this flag and, if set, lists the status from the print buffer on the terminal. If both repeaters are enabled, 2-meter status is printed first.

Each status line printout includes the hour, the number of seconds of total transmission, the number of QSO periods, timeouts and IDs issued. For status purposes, a QSO is defined as a period of exchanges separated by no more than thirty seconds. The last printout column is the number of receptions too short to bring up the repeater.

reference

1. *9900 Family Systems Design*, publication LCC4400, Texas Instruments, Incorporated.

ham radio

6-meter amplifier

A companion unit to the 2-meter and 1 ¼-meter amplifiers

This six-meter amplifier is a companion unit to the 2-meter and 1-1/4-meter amplifiers previously described in *ham radio* articles.[1] All three amplifiers are built using the same chassis configuration originally described by K2RIW for a stripline kilowatt for 432 MHz.[2] The 50-MHz version uses a conventional pi-network output with inductive tuning, and a coil simulated half-wave line for its input section. Both the tetrode (**fig. 1**) and the triode (**fig. 2**) versions will be discussed. Like its predecessors, the 50-MHz amplifier uses parallel combinations of any of the 4CX250 type tetrodes, the 8930 tetrodes, or the 8874 triodes. Metering and power supply connections are identical to the 2-meter and 1-1/4-meter amplifiers. Using a standard design for VHF/UHF amplifiers, a single power supply can be switched from one amplifier to another. Remote operation with a separate metering unit at the operating position or built into the power supply is another adaptation, useful particularly at 432 MHz.

These four 12 × 6 × 8 inch (30.5 × 15.2 × 20.3 cm) power amplifiers for the four popular VHF/UHF

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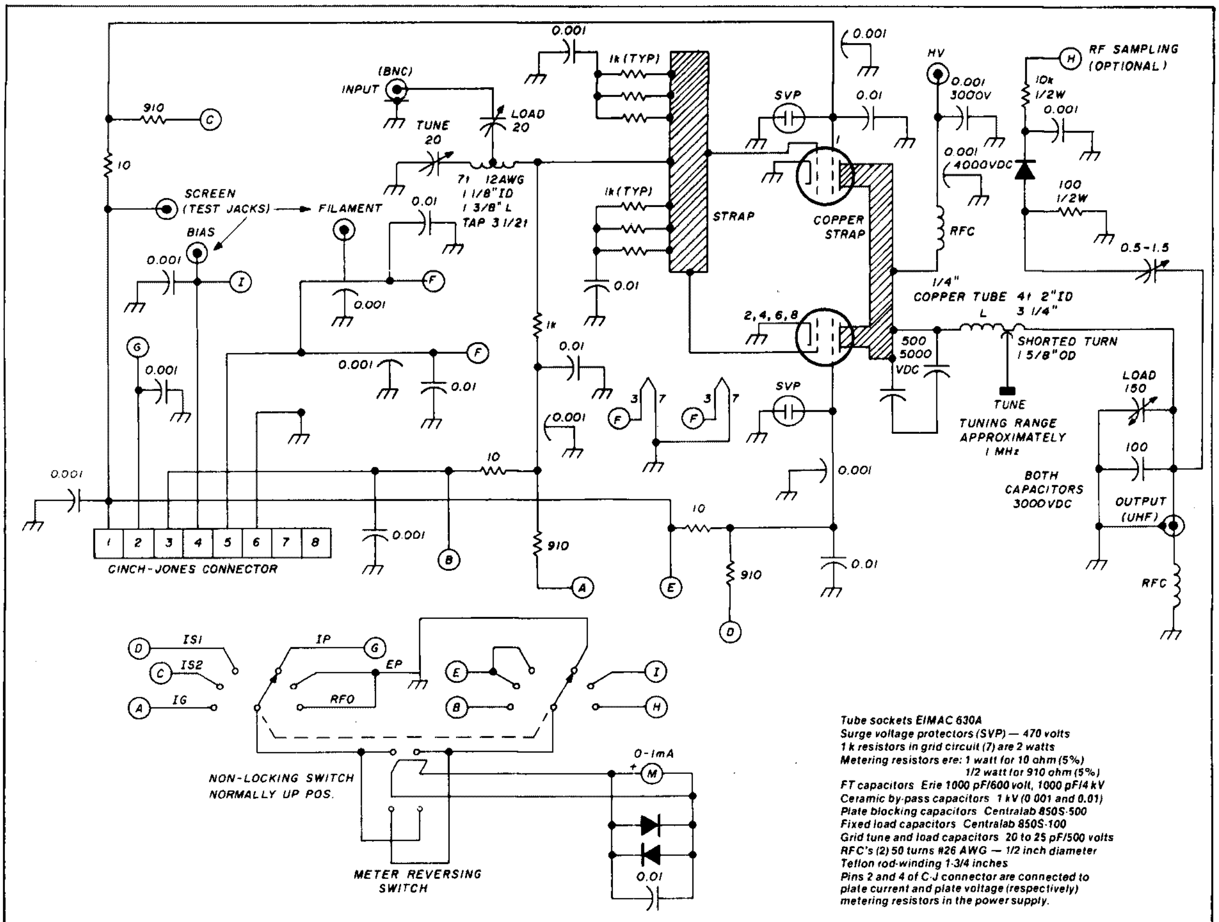


fig. 1. Tetrode amplifier schematic.

Amateur bands have been successfully duplicated hundreds of times. They are rugged and offer a proven performance developed by thousands of hours of testing and use over the past eight years. They provide flexible and reliable high-power operation.

By initially drilling and punching a set of chassis boxes for all four models (432, 220, 144, and 50 MHz), an amplifier can be converted from one band to another. This might be achieved by using a quick-change mechanical procedure for the four separate frequency-sensitive circuit elements.

construction details

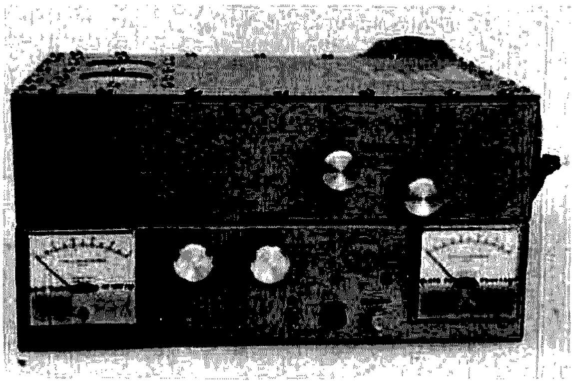
The essential dimensions for chassis drilling and punching are contained in the articles listed in reference 1. This article covers only construction details peculiar to the 50-MHz amplifier.

Referring to the schematic of the 50-MHz tetrode amplifier (**fig. 1**), notice that the two grids are connected by a copper strap between the sockets. The two anodes are paralleled by a brass or copper plate assembly which uses fingerstock for connection to the anodes, providing a mounting for the plate blocking capacitors and a connection point for the high-voltage RFC. The dc circuitry is similar to that found in the previously described amplifiers.

In the triode amplifier (**fig. 2**), the rf section is exactly the same as that shown in **fig. 1** except that rf chokes are used in the filament leads and in the cathode bias lead. The cathode bias and metering circuitry is conventional for a grounded grid amplifier. Two meters are used with the grid current meter on a non-locking switch to read plate voltage.

control and safeguard options

The optional circuitry shown in **fig. 2** provides examples of control and safeguard features which can be added to these amplifiers. The blower option provides 120 Vac on pins 2 and 4 of the cable connector. This permits powering the blower from a receptacle on the amplifier chassis, rather than running a lead



Tetrode amplifier — front view.

back to the power supply. An air switch is mounted in the blower air stream and connected via the blower connector to two power switches (one locking and one non-locking) and to pin 7 of the amplifier connector. Pin 7 is the power relay operate lead in the power supply.[1]

To turn the amplifier on, the locking-type power switch is switched to the on position and the non-locking (push-button type-momentary) switch is pressed to operate the power relay. The power relay energizes the power supply and provides 120 Vac on pins 2 and 4 to start the blower. With the blower up to speed, the air switch keeps the power relay actuated. Once the push button is released, the power supply relay is under the control of the air switch.

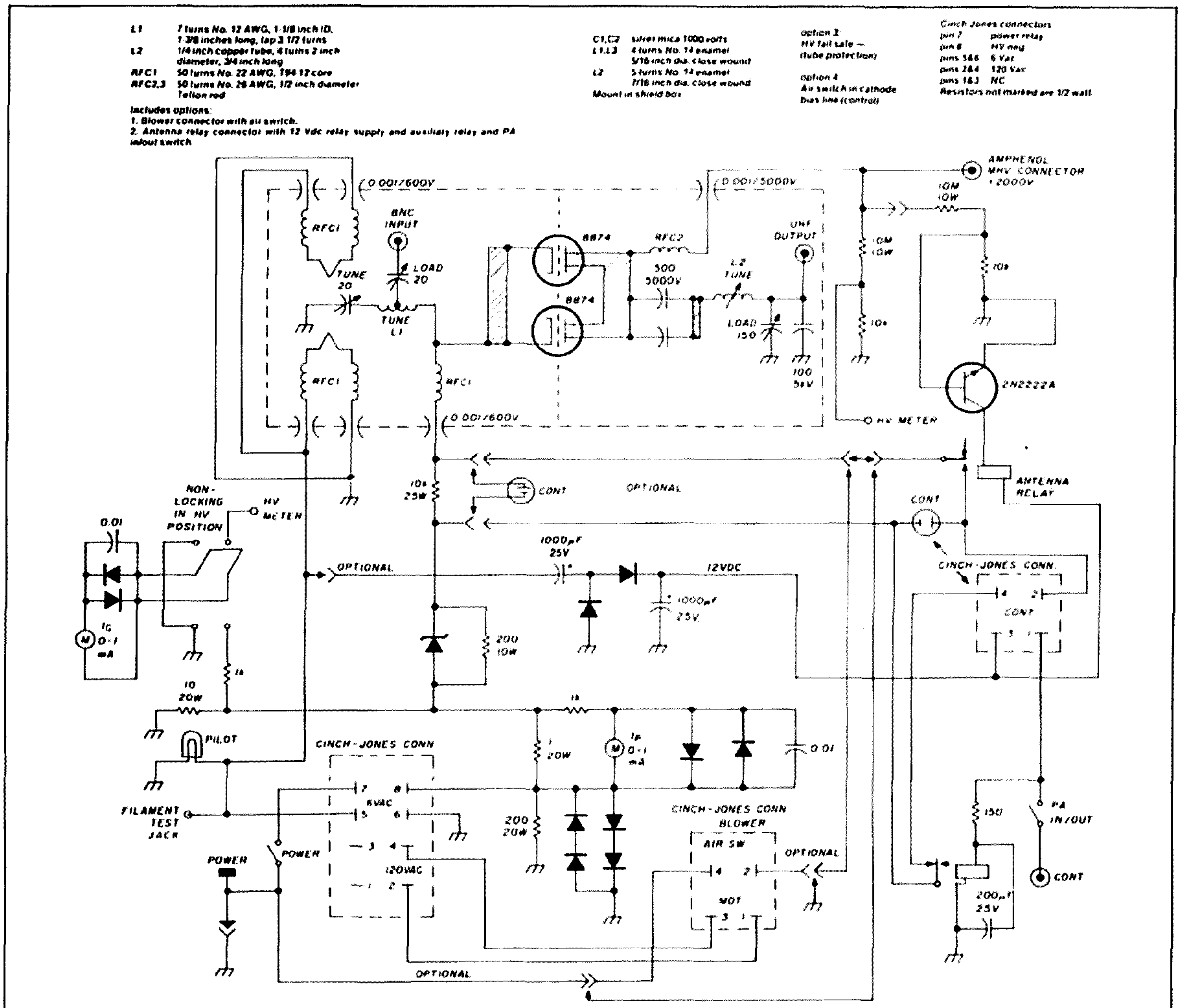
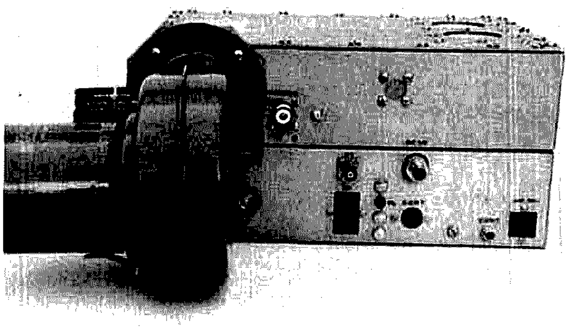
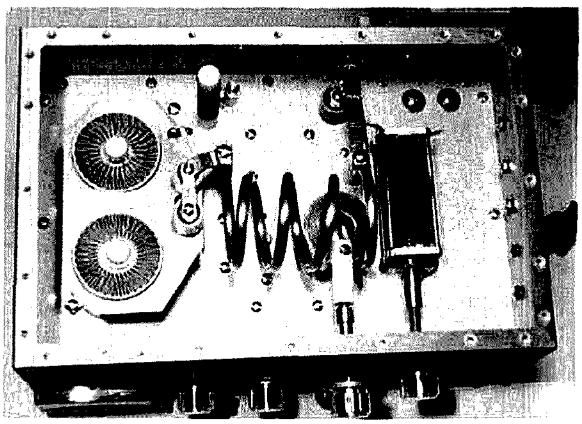


fig. 2. Triode amplifier schematic.



Tetrode amplifier — rear view.



Tetrode amplifier — upper chassis.

Should the blower fail or not come up to speed, the power supply will automatically shut down, an important safeguard considering the two hundred dollar price tag on 8874s.

If excitation is applied with no plate voltage on the tubes, damage to the grid structure may result. The high-voltage fail-safe option provides a safeguard by using a transistor and a relay to open the bias control circuit if high voltage is not present. A 12-volt power supply for this feature is provided by a voltage doubling circuit from the filament line.

The remaining option, shown in **fig. 2**, is used to operate a DPDT coaxial relay which can be mounted (with a coaxial adapter) on the output connector of the amplifier. The coil of the relay and a set of auxiliary make-contacts are connected to the amplifier chassis via a four-contact connector. The 12-volt supply, auxiliary control relay circuitry, a power amplifier (PA) in/out switch, and a control jack complete this feature. Note that a ground on transmit to the amplifier control jack will apply operating bias to the amplifier only if the antenna relay is operated and the auxiliary relay (in this optional circuit) is released. In receive, 12 volts is applied through the winding of the antenna relay to the auxiliary relay winding. The auxiliary relay operates, but the antenna relay, which requires more current than the auxiliary relay, does not operate with the PA switched to the in position. A ground on transmit from the exciter causes the antenna relay to operate immediately and the auxiliary relay to release after a slight delay. This prevents the amplifier from being "hot switched" and provides additional protection for the rf amplifier in the receiver. A layer or two of cellophane tape on the pole piece of the antenna relay is usually required to guarantee release. More sophisticated antenna relay-control circuitry is desirable, however, for EME amplifier applications.

Construction and mounting arrangements for the various options are covered in the construction infor-

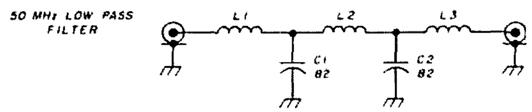
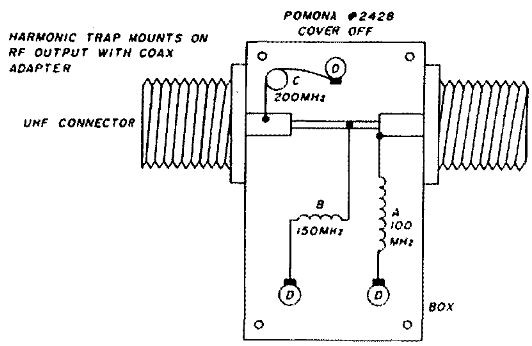
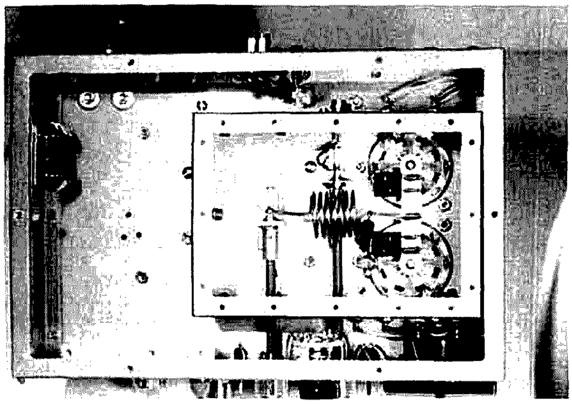


fig. 3A. 50 MHz low-pass filter.[3]



A 20 turns No. 22 AWG, 3/4 inch long 1/4 inch Teflon rod
B 8-1/2 turns No. 20 AWG, 1/2 inch long
C 6 turns No. 20 AWG, 1/2 inch long
D Piston trimmer capacitors 10 pF
50 MHz amplifier
upper chassis — rear
RF output drilling
L P filter

fig. 3B. Trap filter.



Tetrode amplifier – lower chassis.

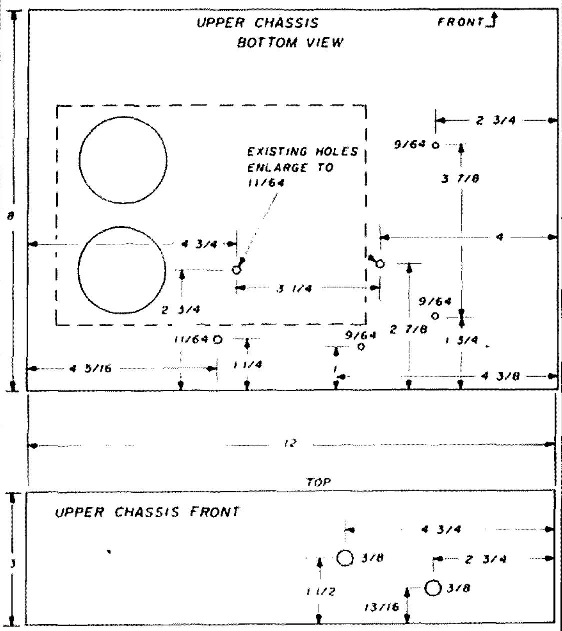


fig. 4. Upper chassis drilling chart.

mation for the triode amplifier. Which options are chosen, and whether they are mounted inside or outside the amplifier, is determined by the intended application and the builder's inclination. These options are also applicable to the 50-MHz *tetrode* amplifier version as well as to the other models of these amplifiers, already described.

A lowpass filter or harmonic trap circuit is needed in the rf output to attenuate harmonics in the amplifier output. These amplifiers, even when operated in the linear mode, may have harmonic components no more than 40 dB down from the fundamental, a level of harmonic attenuation which no longer meets modern RFI design requirements. A suitable LP filter design for this 50-MHz amplifier is shown in the 1981 *ARRL Handbook*, pages 7-11 (fig. 3A). Harmonic trap circuit construction is shown in **fig. 3B**.

Information on the triode and tetrode amplifier power supplies has already been provided in the 220-MHz amplifier article.[1]

construction – tetrode amplifier

If you do *not* intend to use the chassis for the 50-MHz amplifier on any of the other VHF/UHF bands, *omit* the following in its construction: five holes (11/64 inch or 4.4 mm) in the right side of the upper chassis used for mounting the 2-meter plate line, four holes (7/64 inch or 3 mm) and one hole (5/8 inch or 15.9 mm), on the rear of the upper chassis for mounting the rf output connector; two holes (7/64 inch or 3 mm), one hole (3/8-inch or 9.5 mm) for the plate load control in the top plate, and the hole in the front of the lower chassis for the plate tune control. The remaining holes not used for 50 MHz can be drilled and disregarded or filled with 6-32 (M3.5) hardware.

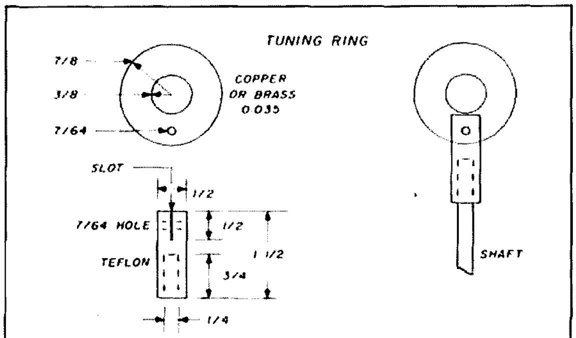
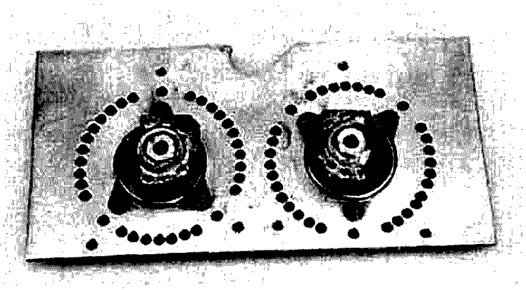


fig. 5. Inductive tuning ring details.



Triode amplifier socket plate assembly – bottom view.

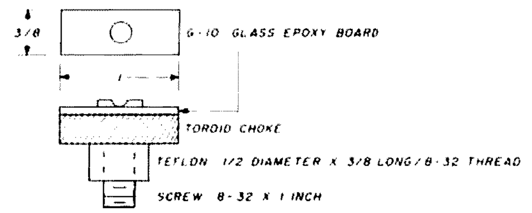
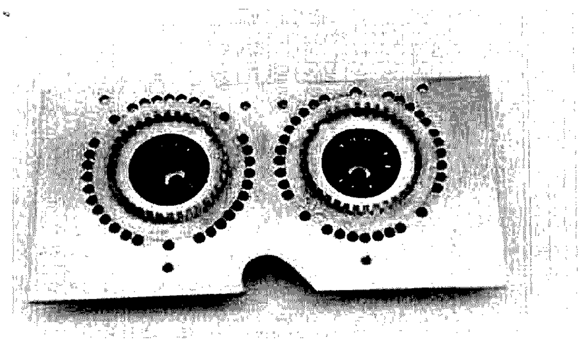


fig. 6. Toroid choke mounting detail.



Triode amplifier socket plate assembly — top view.

Fig. 4 shows the upper chassis drilling required for mounting the plate coil, variable load capacitor, rf choke, fixed load capacitor, and tune and load controls. **Fig. 7** shows the drilling and punching for the rf output connector. This completes the chassis preparation.

Details of the inductive tuning ring are shown in **fig.** 5. **Fig.** 8 gives the dimensions for the plate line. **Fig. 9** provides information on the plate rf choke.

The plate coil is wound with 1/4-inch (6.3-mm) copper tubing, four turns, 2 inches (50 mm) ID, 3-1/4 inches (8.3 cm) long. The ends of this coil are flattened, bent and drilled 11/64 inch (4.4 mm), to mount the coil on 1-1/2-inch (3.8-cm) Teflon pillars midway between the top and bottom of the upper chassis. When construction is completed, the spacing between the turns of the plate coil is adjusted to provide the required tuning range. The tuning range with the inductive ring is in excess of 1 MHz. An accurate grid dip meter is useful for preliminary adjustment of turns spacing for the desired frequency range. The final adjustment of coil size to the desired range is made during the final rf testing.

The assembly and wiring may be done in the same sequence used for the 144- and 220-MHz amplifier, by first assembling and wiring the lower chassis and then assembling the upper chassis and grid box. Mount the sockets and install the plate line parts. Finally, join the upper and lower chassis, make filament and grid bias connections, and install the grid box parts to complete the assembly.

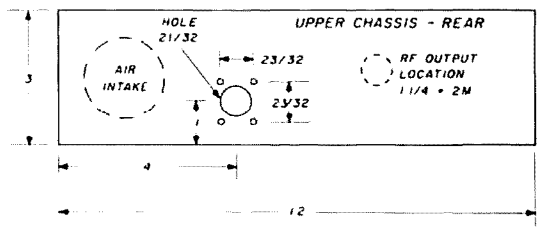
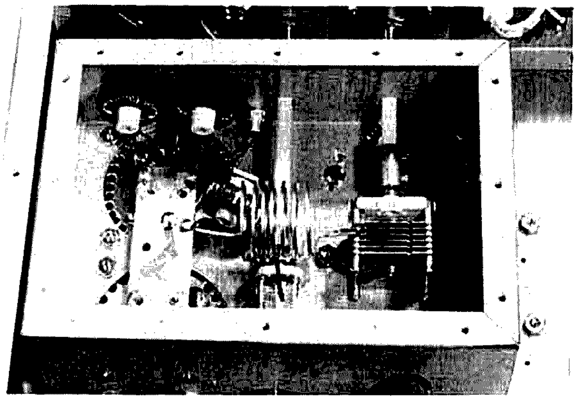
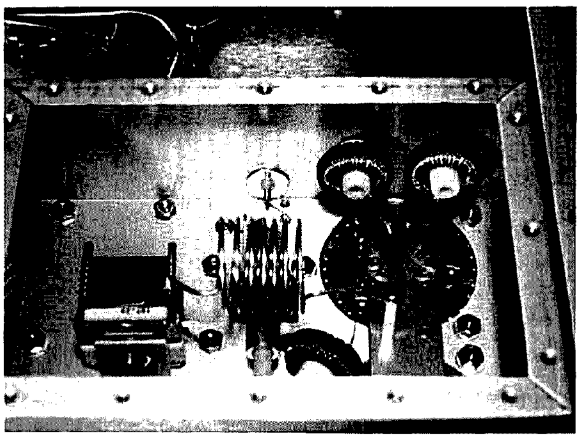


fig. 7. Upper chassis drilling for rf output connector.



Cathode box of triode amplifier viewed from rear of amplifier (toward front of amplifier). Note that toroid choke mountings are not exactly the same as *fig. 6*.



Bottom view of cathode box of triode amplifier.

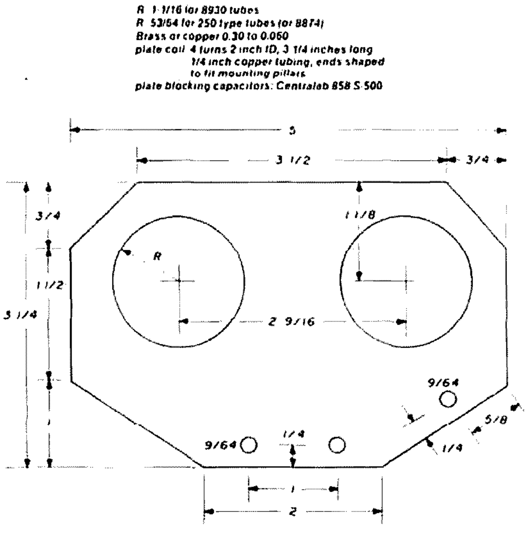


fig. 8. Plate line.

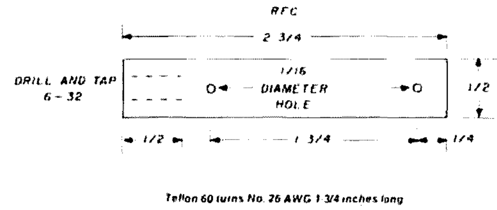


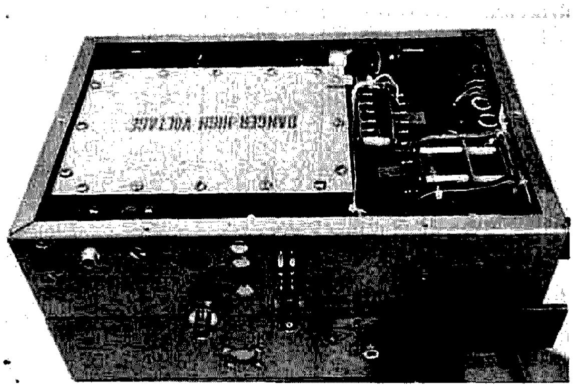
fig. 9. Plate choke.

construction — triode amplifier

Follow the directions for the tetrode amplifier construction for chassis drilling and punching, for the plate line and plate coil. The cathode tuned circuit for the triode amplifier is the same as that described for the grid circuit of the tetrode version. The holes in the grid box for the filament feed-through capacitors are relocated toward the bottom of the box to accommodate the toroid chokes (**fig. 6**). An additional meter hole is punched in the lower chassis front on the right side.

The tube sockets are mounted on a brass plate, as described for the 220-MHz triode amplifier. This assembly (**fig. 10**) lets you solder the grid collet (EIMAC part #882931) in position. Vent holes are provided around the base of the tube; it's a good idea to have this assembly silver plated. The assembly is bolted in place in the same position as the two 630A sockets used for the tetrode amplifier. A brass strip (**fig. 11**) may be used to connect the cathode pins of the two sockets together. This strap is soldered in place after the socket plate has been mounted. Its position is such that the cathode socket pins protrude through the holes about 1/8 inch (3 mm).

Alternatively, a small brass plate mounting a brass bushing (tapped 10-32) may be soldered to the cathode pins of each socket. This method of construction is more involved, but avoids soldering the grid strap in place after the socket plate is mounted. The grid strap is fastened by the 10-32 screws on each mounting plate.



Triode amplifier bottom view to illustrate mounting of optional circuit features on terminal boards in lower chassis.

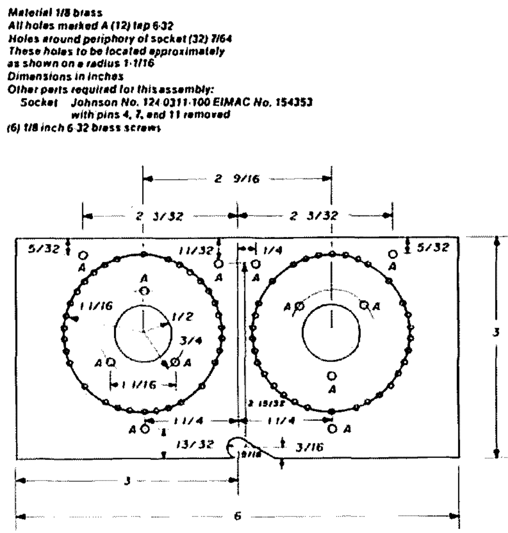
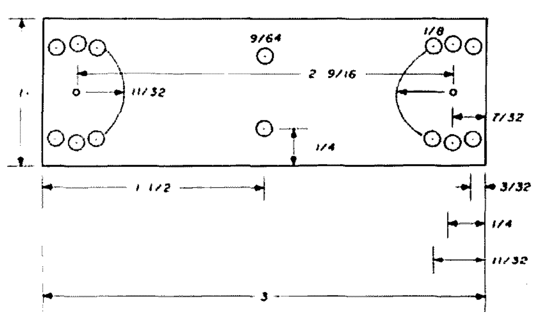
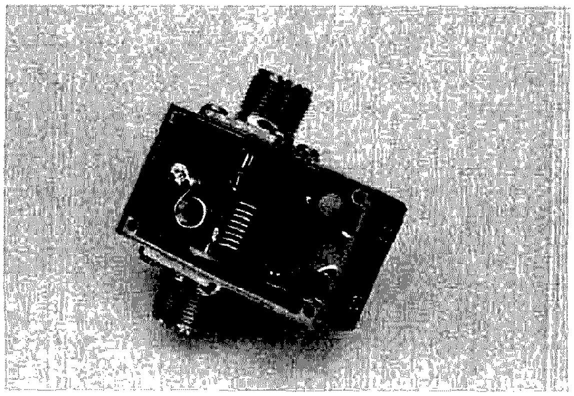


fig. 10. Triode socket mount.



Grid collet for 8874 EIMAC No. 882931
Strap between cathode pins 0.030 brass

fig. 11. Cathode strap.



Output harmonic trap assembly. Three series traps — 100 MHz, 150 MHz, and 200 MHz. Piston capacitor adjustment screws (3) protrude out the bottom of the box.

Metering and other circuitry is mounted in the lower chassis, as shown in the photos. The vitreous-type resistors are mounted to the chassis wall. Other resistors and parts are mounted on terminal boards secured to the chassis with mounting spacers.

The options shown on the triode amplifier schematic (fig. 2) are mounted as follows:

The antenna relay connector is located on the right side of the lower chassis (rear). The small relay associated with this option is located in any convenient spot in the lower chassis. The various resistors, capacitors, and other parts for the antenna relay control circuit, the 12-Vdc supply, and the high voltage fail-safe circuitry are on terminal strips which are located in the lower chassis.

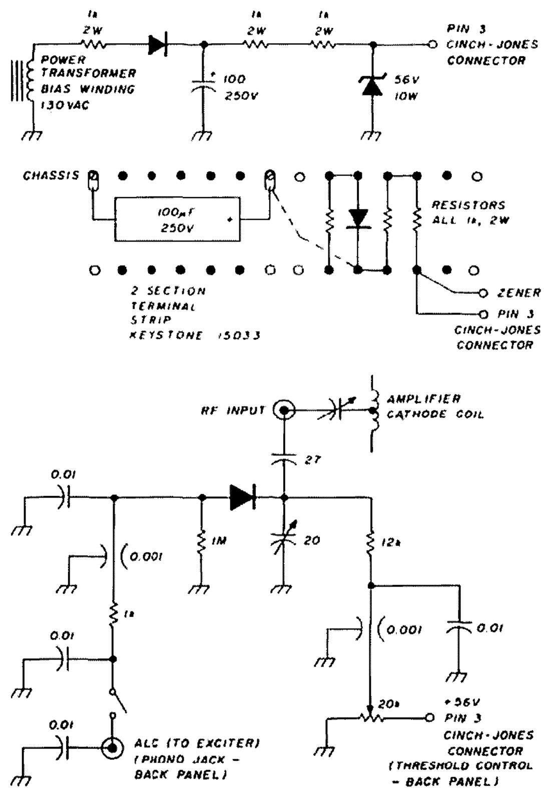
The blower connector is located on the left (side) rear of the lower chassis.

The PA in/out switch, the power switch, and the non-locking switch to start the blower are located on the front of the lower chassis.

In assembling and wiring the triode amplifier, follow the same pattern described for the tetrode amplifier — lower chassis parts mounting and wiring first — upper chassis and cathode box, tube socket assembly, plate circuit parts, joining upper and lower chassis, cathode parts, and the final wiring steps.

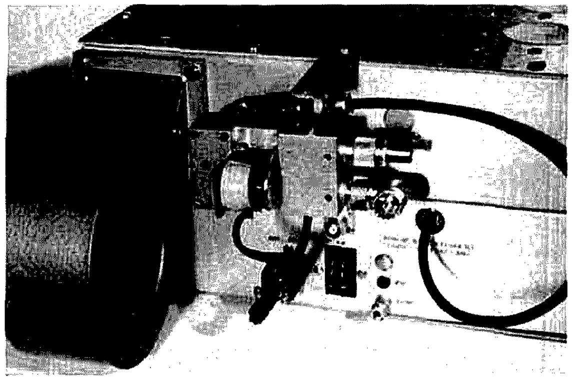
automatic load control

An ALC circuit (**fig. 12**) has been added as an option to the triode amplifier. The parts within the grid box are mounted close to the rf input connector. A bias winding is required on the high-voltage transformer, or a separate small transformer is required to provide the +56 volts threshold control voltage. The bias voltage parts can be mounted in the power supply chassis on a terminal board.



Note: The circuitry above is installed in the high-voltage power supply. As indicated, the 56 Vdc is connected to the ALC circuitry in the amplifier via Pin 3 of the low-voltage cable.
Note: The circuitry up to the 0.001 FT capacitors is installed in the cathode box of the amplifier. The balance of the circuitry is in the low chassis.
ALC control option
50 MHz 8874 amplifier
W2GN 02-20-81

fig. 12. ALC circuit.



Mounting of double-pole coaxial relay on output connector.

table 1. Typical operation tetrode amplifier.

drive power	grid current	screen current	plate current	plate voltage	power output
0	0	0	0.100	2150	0
2.5	0	0	0.260	2010	177
5.0	0	−0.003	0.430	2000	470
10.0	0.002	0.027	0.600	2000	800

filament volts = 6.07 grid volts = 64 screen volts = 315

table 2. Typical operation triode amplifier.

drive power	grid current	plate current	plate voltage	power output
0	0	0.040	2300	0
2.5	0.002	0.210	2100	140
5.0	0.004	0.300	2050	285
10.0	0.025	0.380	2050	540

filament volts = 6.12

Note: The triode amplifier may be driven to an output level of 1 kW (SSB).

operation

The 50-MHz amplifiers tune and load in a conventional manner. Make initial adjustments with low drive power. Final adjustment of the grid (or cathode) tuning is made for lowest SWR toward the drive source. Final adjustment of the plate tuning must be done at full power output in order that the load control may be set at its optimum position.

Tables 1 and 2 show typical operation of the tetrode and triode amplifiers.

references

1. Fred Merry, W2GN, "Stripline Kilowatt For Two Meters," *ham radio*, October, 1977. Also, "Stripline Kilowatt for 220 MHz," *ham radio*, April, 1982.
2. Richard T. Knadle, Jr., K2RIW, "A Stripline Kilowatt for 432 MHz," *QST*, April, 1972, page 48; May, 1972, page 59.
3. *ARRL Handbook*, 1981, pp 7-11.

ham radio

the ham notebook

FSK tone generator using an integrated tone dialer

Have you ever thought about redesigning or building an FSK (frequency shift keying) tone generator? If so, you are not alone. How many FSK

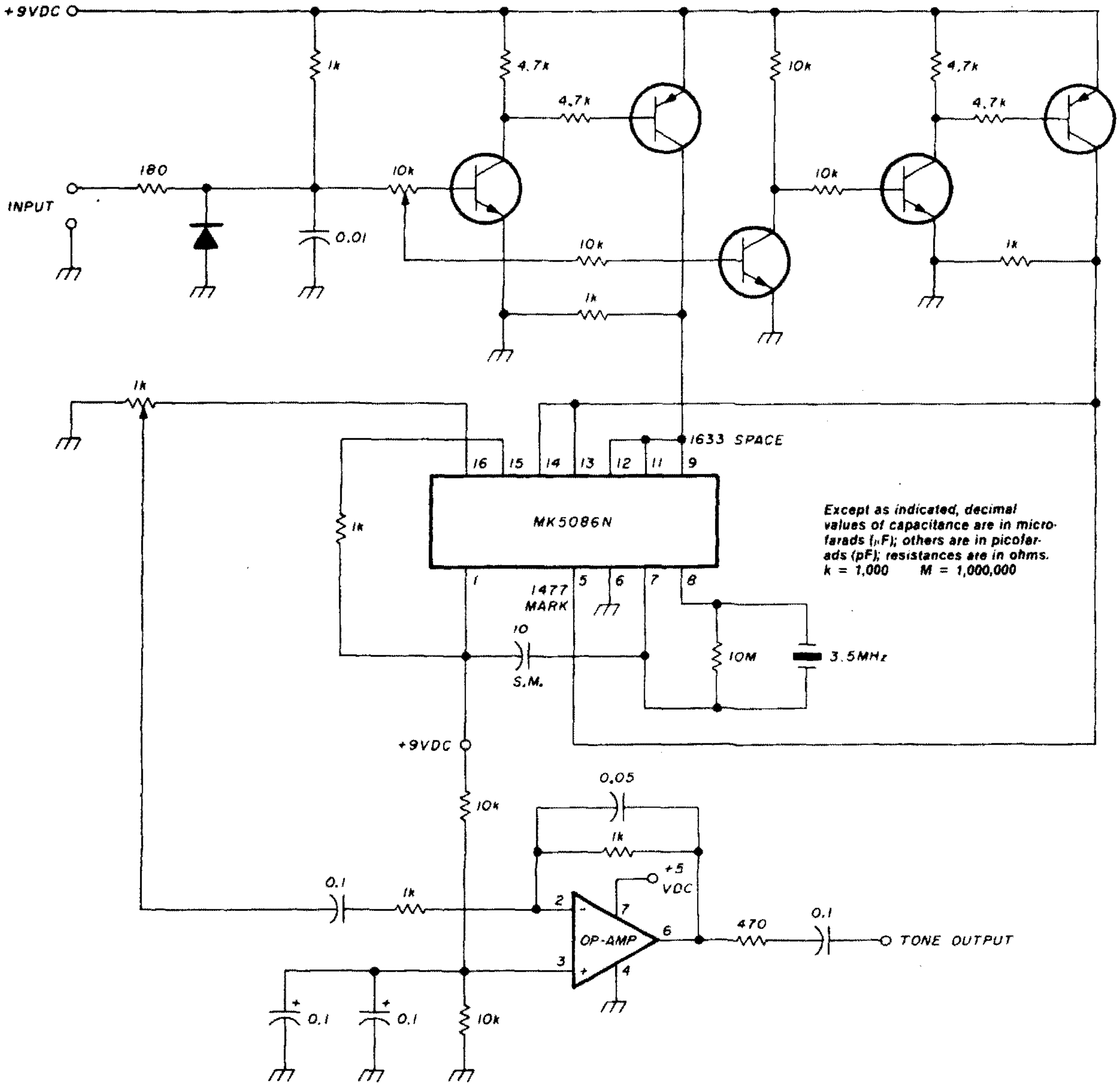


fig. 1. Schematic for the FSK tone generator.

generators have you seen that use an integrated circuit? To my knowledge there aren't very many.

Here is an easy way to build a circuit using a TouchTone™ chip which generates the frequencies needed for FSK. This circuit is connected between the teleprinter and transmitter. There are four main areas in constructing this circuit: the integrated tone dialer chip, switching circuit, filter, and amplifier. See **fig. 1**.

frequencies

Both frequencies are generated by a Mostek MK 5086N IC chip. Pin 9 is used for space at 1.633 kHz, and pin 5 as mark at 1.47 kHz. A 3.579545 MHz television color-burst crystal is the frequency-determining element for the chip. To simulate keyboard operation, tie pins 14 and 13 to pin 5 and pins 12 and 11 to pin 9. This makes the Mostek think it is being switched by a keyboard.

Transistors in a switching circuit determine if a space or a mark is sent.

filter and amplifier

An op amp provides a small amount of needed gain. A lowpass filter is used to reduce the harmonic content generated by the Mostek IC chip. This filter can be made by placing a capacitor across pins 2 and 6 of the op amp.

This circuit was constructed by Charles Aron, Ney Vew, and David Nagel at Northern Montana College in Havre, Montana. Special thanks are also given to Lee Barrett; without his time and advice this project would not have been possible.

David Nagel
Havre, Montana

capacitive-reactance meter multiplier

Recently I saw a large commercial type 0-150 Vac voltmeter in mint condition — just what I needed for my station control panel to monitor line voltage. However, the external series resistance was missing. Well, the owner sold it to me for $2.50, as he admitted it didn't have too much value as it was. I discovered it would need an external 15-watt series resistance of about 1500 ohms. I decided to use a capacitor of the same reactance instead of using a resistance; reactances do not dissipate power and I would save energy.

The calculation for finding the required reactance is:

$$C = 1,000,000/(6.28)(f)(X_C)$$

where f is the line frequency, in this case 60 Hz, X_C is the desired reactance in ohms equal to 1500 ohms, and C is the required capacity in μF:

$$C = 1,000,000/(6.28)(60)(1500)$$
$$= 1.77\,\mu F$$

The theory and application worked fine. I used a good accurate ac voltmeter as my calibration standard. By paralleling a few small non-electrolytic fixed condensers from my junk box, it was easy to make my meter read the same. The real advantage of using condensers is that the power drain on the line is practically negligible. Naturally, the calibration is good only for the 60 Hz line voltage you are monitoring.

William Vissers, K4KI

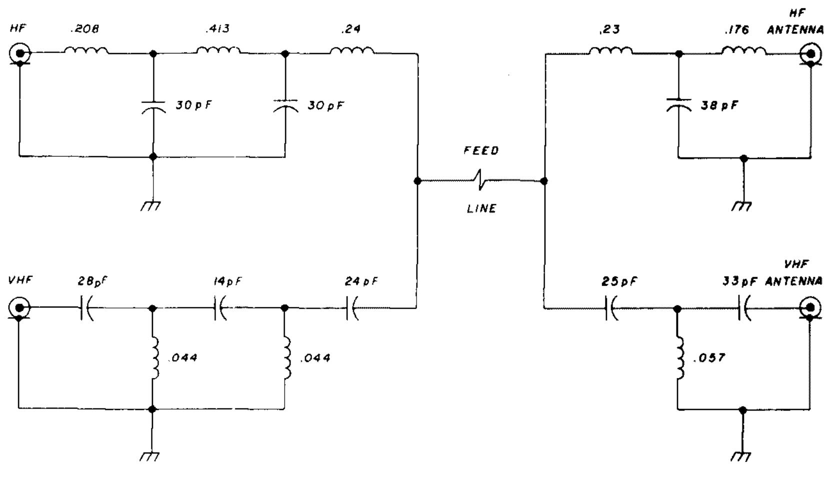


fig. 2. The capacitor changes to the N6RY diplexer mods.

diplexer mods

You can diplex high frequency to go above 28 MHz (refer to N6RY's article on page 71 in the December, 1980, issue of *ham radio*). By building up the VHF part of the two boxes and changing a couple of the capacitors in the high frequency side of the boxes, you can operate 10 and 2 or 10 and 220, or 6 and 2 or 6 and 220 meters at the same time. You can also add 6 or 10 meters to your 2 or 220 repeater by adding a box and an antenna on top and a box and a repeater or remote base on the bottom. All additions use the same feed line. The capacitor changes are shown in **fig. 2**.

If you have a 6-meter rig and want to go mobile, but can't find a spot to mount another antenna, try a 5/8 wave 2-meter antenna and check the SWR. If it is low on 6, just add the box between the 6 and 2-meter rigs and connect it to the same antenna.

Robert McWhorter, K5PFE

simple diode tester

I recently had to check the peak inverse voltage of some surplus diode units. Searching for a suitable device, I decided to use a high-resistance transformer acquired at a flea market sale. This particular unit had a high resistance secondary (over 600 ohms) which precluded its use for service in a power supply unit supplying more than minimal power. This was hooked up as shown (**fig. 3**), in a simple full-wave doubler circuit, and provides over 1,000 volts dc from a secondary rated 400 volts ac.

There are two methods for checking diodes for PIV. One method is to increase the test voltage until there is 10 μA of reverse current (for a 1-ampere diode) and then to rate the diode at a safe peak inverse voltage of 20 percent lower. The method I prefer is to calibrate for a PIV of that value attained when 5 μA of reverse current flows. Either way gives a satisfactory rating for diode breakdown voltage, see **fig. 4**.

Any multimeter with a basic sensitivity of at least 5,000 ohms per volt can provide the needed test current, since the basic limiting resistance is present in the meter's multiplier resistance. A convenient method of checking voltage at the same setting is to simply short out — with an insulated screwdriver — the terminals across the diode being tested. The high-resistance secondary precluded the need for any limiting resistors in the circuit, and the low-capacity filter capacitors cause the output voltage to drop sharply under load, tremendously reducing the hazard of testing with high voltage sources.

Neil Johnson, W2OLU

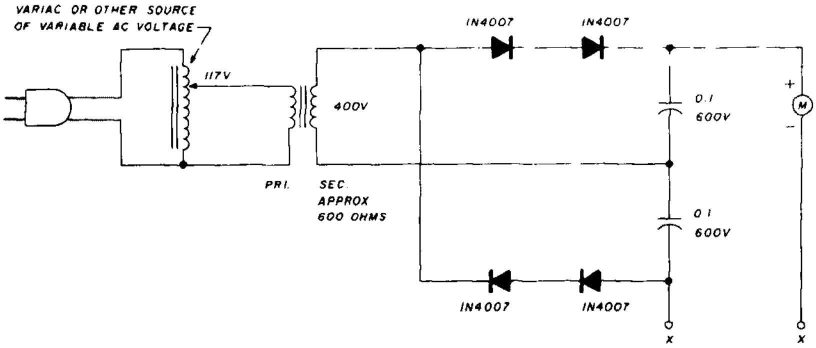


fig. 3. Simple tester for checking silicon diodes. To test diode, insert in circuit at x, and then raise input voltage from zero. Meter M is any sensitive voltmeter on 1,000 volt scale, having sensitivity of 5,000 ohms per volt or more. Alternate method is to utilize a 0-200 microammeter and 5 megohms of resistance.

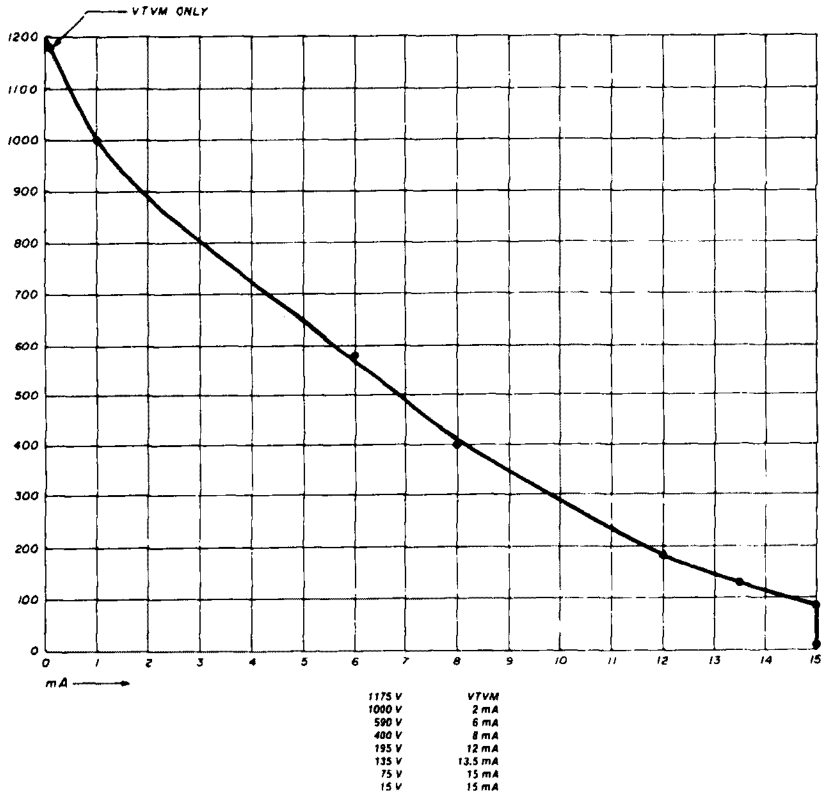


fig. 4. Sample rating chart. Safety is enhanced by limiting current by high impedance supply.

improved logic probe

I was considering buying or building a logic probe to complement my dual channel scope when troubleshooting my homemade microcomputer. The August, 1980, issue of *ham radio* finally convinced me to build my own version.

The following specs were essential: indication of high-low-open conditions; capture-stretching short-positive or negative pulses; operation at TTL (5V) and at CMOS (5-15V) levels; high and low should be indicated at the specified levels for each logic family and every voltage, that is, 0.8 and 2.5V for TTL and 1/3 V_{cc} and 2/3 V_{cc} for CMOS.

I took two ideas from N6UE's article[1] in the August issue on page 38: using the 555 timer and voltage regulation for the display LEDs[2] I met the requirements of the first, third, and last specs by using National Semiconductors' 339 quad, single supply comparator[3] I obtained the required reference levels from a voltage dividing network and a switch, which modifies the resistor values to suit TTL-CMOS levels. See **fig. 5**.

Comparators a and b serve as a window detector, both being high inside the forbidden voltage region, while going low at a high or low input. A low from comparators a and b is used to drive the high (red), and the low (green) LEDs. The negative transitions are differentiated and ORed by the remaining two comparators, and applied to the 555 for stretching. The timer drives the pulse (orange) LED. An LM309 TO-5 voltage regulator provides protection for the LEDs against voltage rise.

I wired the prototype on a piece of Veroboard. As I lack a PC board production capability at home, I decided to stay with the prototype.

Tests indicate that the probe operates as required up to about 250 kHz square wave input. The minimum captured pulse width is about 4 μs. These results are close enough to the specified delay through the comparators to indicate that speed-pulse width limitations could be reduced by using faster comparators.

references

1. R.S. Isenson, N6UE, "Digital Logic Probe," *ham radio*, August, 1980, page 38.
2. Signetics NE555V data sheet.
3. National Semiconductors LM339A data sheet.

J. Rozenthal

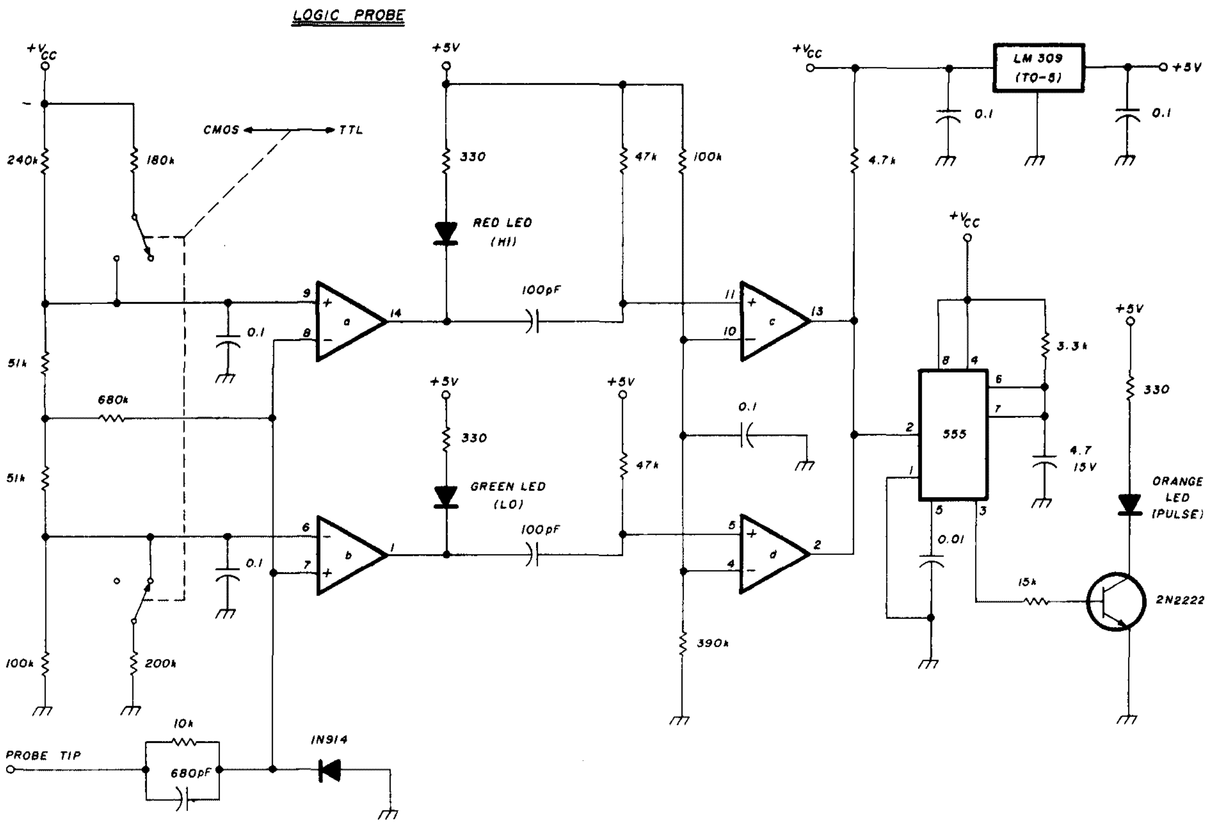


fig. 5. The improved logic probe.

Garth Stonehocker, K0RYW

DX FORECASTER

spring DX

The powerful DX months (around the equinox) are here for us to try again. Over the years March and April have provided excellent 6-meter openings on transequatorial (TE) paths. Using 6-meter openings as a criterion for the higher-frequency-band DX, last year didn't have as many openings as 1981, but the opening on March 4 was acclaimed the best in years in *Westlink Report*. The March 4 opening was a period of high solar flux and geomagnetic disturbance, which probably influenced the TE (one-long-hop) propagation. April was also roaring with TE openings from the southern U.S. to South Africa, South America, New Zealand, and Australia. The other openings in April were not so pronounced, as the solar flux was lower. However, four large disturbances (April 2, 11, 25, and 29) and two smaller disturbances (April 17 and 21) increased the ionization near the geomagnetic equator for high maximum-usable frequencies for TE.

This year's 6-meter openings may be fewer in number since we are already near the half-way point on the down-side of cycle 21. The sunspot number should be about 75 (123 flux units). The second maximum 1981-82-83) period of geomagnetic-ionospheric disturbance in cycle 21 is expected to be the dominating factor for openings this year. These disturbed periods during April are expected around the 5th, 15th, and 23rd. The latter is the longer recurrent type (see February, 1982 DX Forecaster).

last minute forecast

The higher segment of the h-f bands (6-30 meters) will probably be best during the middle of the month. Watch for the high radio flux and disturbance numbers from WWV at 18 minutes after the hour. On the lower bands (30-160 meters)* night DX will be best during the first and last weeks of the month, particularly in-between the springtime frontal thunderstorms when QRN should be low. Your favorite TV weather forecaster will show these fronts moving across your QTH.

The perigee of the moon's orbit (for moon-bounce DX) is on the 21st at 2100 hours; the moon will be at full phase on the 27th at 0631 hours. There will be a short meteor shower, the Lyrid, on April 20-22. The rate is five per hour, hardly a real help for *meteor-scatter* DX. But a bigger shower, the Aquarid, starts before the end of the month, peaks on May 5, and ends by mid-May. Its rate is 10 to 30 per hour.

band-by-band forecast

Six meters may provide occasional band openings with a peak during the late afternoon hours. Transequatorial north/south paths will be the best. Your guide to good conditions are strong openings on 10 meters with high values of solar flux and A and K geomagnetic indices.

Ten and fifteen meters will be open to many areas of the world from morning until early evening hours most days. Times of geomagnetic disturbances will limit the number of signals heard, but listen carefully — they can be from very unusual places. Fifteen meters should stay open later in the day than 10 meters. Operate 10 first and move down to 15. More hours of daylight means earlier band openings and longer periods of operation.

Twenty meters will be the main daytime DX bands, as it is almost always open to some part of the world. It opens to the east as the sun rises and extends into the late evening hours to the west. Geomagnetic disturbances do not affect this band as much as the higher ones, but look for unusual transequatorial DX propagation once in a while. One-hop transequatorial DX of 5,000 to 7,000 miles (8,000 to 11,200 km) may be possible in the late evening hours during some of these unusual conditions.

Thirty meters is a day and night band. The day portion should be like 20 meters except the signal strengths may decrease during midday on some days. Days of decreasing strength should be those with high solar flux values. This band will also work well into the night, often through the night. Nights this doesn't hold true will most likely follow a day with a very high solar flux value. The problem time is usually the hour or so before dawn. The workable distance may be expected to be greater than 80 DX at night and less than 20 during the day.

Forty and eighty meters will exhibit short skip conditions during daylight hours and lengthen after dark. The bands will open to the east just before your sunset, swing more to the south toward Latin America about midnight, and end up in Pacific areas during the hour or so before dawn. On some nights these bands will be as good as during the winter DX season. The coastal regions usually have the edge for working rare DX on these bands.

One-sixty meters will probably bring many nights that will remind you of last summer's noise. However, many good nights are left for working DX before this summer's noise comes to stay. Propagation on 160 meters will approximate a shortened 80-meter condition.

ham radio

*Editor's note: 30 meters because of its unique place in the h-f spectrum and characteristics is discussed in both sections (higher/lower segments) of the h-f band forecast.

APRIL

WESTERN USA

GMT	PST	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	4:00	15	20	15	10	15	10	10	10
0100	5:00	15	20	20	10	15	10	10	10
0200	6:00	15*	20	20	10	15	10	10	15
0300	7:00	15	20	20	15	15	10	10	15
0400	8:00	15	20	20	15	15	10	10	15
0500	9:00	20	20	30	15	20	10	15	15
0600	10:00	—	40	30	20	20	15	15	20
0700	11:00	—	40	20	20	20	15	15	20
0800	12:00	—	—	20	20	20	15	15	20
0900	1:00	—	—	—	20	20	15	20	20
1000	2:00	—	—	—	—	20	20	20	20
1100	3:00	—	—	—	—	30	20	20	—
1200	4:00	—	—	—	—	—	20	20	—
1300	5:00	—	—	—	20	—	—	20	—
1400	6:00	—	—	20	20	—	—	—	—
1500	7:00	—	—	20	15	—	—	—	—
1600	8:00	—	20	15	15*	—	15	—	—
1700	9:00	20	20	10	15	—	15	15	—
1800	10:00	15*	20*	10	10	—	15	15	—
1900	11:00	15	20*	10	10	20	15	15	—
2000	12:00	15	20*	15	10	15	10	15	15
2100	1:00	15	20*	15	10	15	10	10	15
2200	2:00	15	20*	15	10	15	10	10	10
2300	3:00	15	20	15	10	15	10	10	10

MID USA

GMT	MST	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)	CST
0000	5:00	15	20	15	10	15	10	10	10	6:00
0100	6:00	15	20	15	10	15	10	10	10	7:00
0200	7:00	15	20	15	10	15	10	10	15	8:00
0300	8:00	15	20	20	15	20*	10	15	15	9:00
0400	9:00	20	20	30	15	20*	15	15	15	10:00
0500	10:00	—	20	30	15	20	15	15	20	11:00
0600	11:00	—	20	30	20	20	15	20	20	12:00
0700	12:00	—	20	20	20	20	20	20	20	1:00
0800	1:00	—	20	20	20	20	20	20	20	2:00
0900	2:00	—	20	20	20	20	20	20	—	3:00
1000	3:00	—	—	—	—	20	20	—	—	4:00
1100	4:00	—	—	—	—	—	—	—	—	5:00
1200	5:00	—	—	—	20	—	—	—	—	6:00
1300	6:00	20	—	—	15	—	—	—	—	7:00
1400	7:00	20	20	20	15	—	—	—	—	8:00
1500	8:00	20	20	15	15*	—	—	—	—	9:00
1600	9:00	20	15	15	10	—	—	—	—	10:00
1700	10:00	20	15	10	10	—	15	15	—	11:00
1800	11:00	20	15	10	10	—	15	15	—	12:00
1900	12:00	—	15	10	10	—	15	15	—	1:00
2000	1:00	—	15	10	10	15	10	15	15	2:00
2100	2:00	—	15	15	10	15	10	10	15	3:00
2200	3:00	—	15	15	10	15*	10	10	15	4:00
2300	4:00	—	15	15	10	15	10	10	10	5:00

EASTERN USA

GMT	EST	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (Caribbean S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	7:00	15	20	15	10	20	10	10	10
0100	8:00	15	20	15	10	20	10	10	10
0200	9:00	15	20	20	15	20	10	15	15
0300	10:00	15	40	20	15	20	15	15	15
0400	11:00	20	40	30	15	20	15	15	20*
0500	12:00	20	40	30	15	20	15	20	20
0600	1:00	20	20	15	15	20	20	20	20
0700	2:00	—	20	15	20	20	20	20	20
0800	3:00	—	20	20	20	20	20	20	20
0900	4:00	—	—	—	20	20	20	—	—
1000	5:00	—	—	—	20	—	—	—	—
1100	6:00	—	—	—	15	—	—	—	—
1200	7:00	—	—	—	—	—	—	—	—
1300	8:00	—	15	—	—	—	—	—	—
1400	9:00	20	20*	15	—	—	—	—	—
1500	10:00	20	15	15	10	—	—	—	—
1600	11:00	15	15	10	10	—	—	15	—
1700	12:00	20	15	10	10	—	20	15	—
1800	1:00	—	15	10	10	15	20	15	—
1900	2:00	—	15	10	10	15	15	15	—
2000	3:00	—	15	10	10	15	10	10	20*
2100	4:00	—	15	10	10	15	10	10	15
2200	5:00	20	20	15	10	15	10	10	15
2300	6:00	20	20	15	10	15	10	10	10

*Look at next higher band for possible openings.

NEW products

REVIEW: Daiwa CN520 SWR and power meter

About the only time we had a chance to use a dual needle SWR meter was when we used the TMC transmitter at W1AW. That was, until MCM (distributors for selected Daiwa products), sent us their CN520 SWR and power meter.

The Daiwa meter comes in four different models: the CN150 for 1.8-60 MHz 20/200 watts, the CN520 for 1.8-60 MHz 200/2 kW, the CN530 for 50-150 MHz 20/200 watts, and the CN540 for 140-250 MHz 20/200 watts. Each of these units measures just 2.83 × 2.83 × 3.62 inches (72 × 72 × 96 mm) and weighs less than a pound. Rf connectors on each are SO-239 and accuracy is listed at ± 10 percent.

Installing the SWR bridge is a matter of connecting it in-line between your transmitter and load. Setting the meter to the correct power position ensures that you will get an accurate SWR reading. Two needles are used to measure SWR: the left needle measures forward power, the right needle measures reflected power. The point at which the two needles cross is the SWR reading. SWR is clearly marked on the meter face by a series of red lines. This is quite handy and allows the operator to know instantly how well his line is matched between transmitter and antenna.

The meter case has two brackets on the side for possible use as a mobile or remote mount.

This is a nice meter. When compared to a lab-type meter, its accuracy is quite good, well within the rated specifications. We find the CN520 to be a breeze to use and a very valuable addition to our ham shack. In fact, after using the dual reading meter it is very hard to use any other kind of unit.

Price is $69.95 retail. For more information, contact your local dealer or MCM direct at 858 E. Congress Park Drive, Centerville, Ohio 45449. Reader Service Number 301.

N1ACH

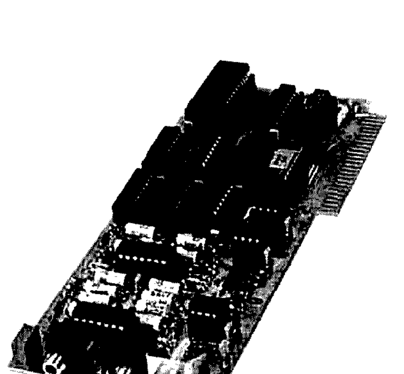
slow-scan TV system

Commsoft has developed PhotoCaster™, a slow-scan television system for the Apple II computer. PhotoCaster provides an easy way for hams who own Apple computers to get started on SSTV with a full-featured black and white and color system. PhotoCaster includes a circuit board to interface an Apple to a TV camera and a receiver/transmitter, plus a two-disk software package.

PhotoCaster can also add titles and graphics, create video special effects, enhance images, retrieve and store pictures on disk, and print high resolution pictures with an MX-80 printer.

Black and white pictures are processed with a resolution of 128 by 128 pixels and sixteen levels of gray. In the color mode, eight colors are available with sixteen saturation levels. Color pictures are taken with an unmodified black and white TV camera using a three-frame RGB sequence. Standard RGB transmission formats are available in addition to a unique Apple-to-Apple single-frame color mode which takes eight instead of the usual twenty-four (or more) seconds to transmit a color picture.

PhotoCaster requires an Apple II or Apple II Plus computer with 48K RAM and one disk drive. The price of the PhotoCaster is $499.95 for the basic system, including an assembled and tested circuit board and software. A complete system, consisting of a Panasonic WV1400 camera, board and software, is available for $749.95.



For more information, contact Commsoft, Inc., 665 Maybell Avenue, Palo Alto, California 94306. Reader Service Number 302.

radio teletype and CW

With the Super-Ratt radio teletype and CW program for the Apple II, you can have your own Radio Bulletin Board System (RBBS) station on-line quickly and easily.

The program will operate in ASCII as well as Baudot at any speed from 40 to 300 baud. CW speeds range from 5 to 100 WPM, with an automatic speed adjust on receive.

The program may be run in either manual or RBBS modes. Extensive use of disk files permits storage of canned material for manual operation. In the RBBS mode, the system automatically saves nearly one hundred user messages to the disk. There are thirty-five different, simple English word commands on the RBBS.

Almost any modern terminal unit or converter can be used with Super-Ratt, as well as devices such as the RADCOM card by AF6W. The program is not protected against copying. The BASIC portion may be listed and modified to suit your tastes. (The registered owner's call is installed in the machine code by the factory.)

A free one-year subscription to the user newsletter, *The Ratt's Nest*, is included in the purchase price of $54.95. For more information, contact Universal Software Systems, Inc., 9 Shields Lane, Ridgefield, Connecticut 06877. Reader Service Number 303.

helical resonator amplifiers

Hamtronics, Inc., has developed a new line of low-noise receiver preamps with helical resonator filters built in. The HRA-144, HRA-220, and HRA-432 units cover the three major VHF and UHF ham bands. The combination of a low-noise amplifier and the sharp selectivity of a three or four section helical resonator increases receiver sensitivity and reduces cross-band interference. The unit has a low 0.6 to 0.95 dB noise figure and 50 to 60 dB rejection of any signals out of the ham band.

The amplifier circuit uses some of the new microwave transistors developed for satellite TV service. Nominal gain is 26 dB on 2 meters, 22 dB on 220 MHz, and 16 dB on 420-450 MHz. A three-section helical resonator is used in the output circuit of the VHF units, a four-section resonator is used in the UHF unit. The VHF unit is only 1½ × 3 inches, and the UHF unit is only 2½ × 3 inches.

The HRA-144 or HRA-220 costs $49.95, and the HRA-432 is $54.95.

For further information contact Hamtronics, Inc., 65 Moul Road, Hilton, New York 14468-9535. Reader Service Number 304.

Ameco multimeters

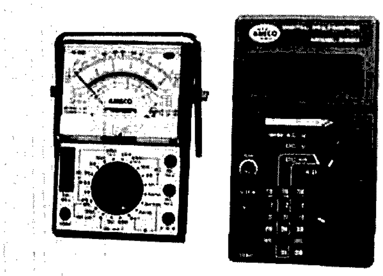
Ameco Equipment Company announces preliminary specifications of its new line of Ameco multimeters. Multimeter Model M-300 (available immediately) features highly sensitive 20K ohms/Vdc and 10K ohms/Vac; gold-

plated switching contacts; overload protection by diodes and fuse; and carrying handle that can be used as adjustable stand.

Ranges for dc voltage: 0-0.25, 1, 2.5, 10, 25, 100, 250, 1000 V; ac voltage: 0-10, 25, 100, 250, 1000 V; dc current: 0-50, 500 μA, 5, 50, 500 mA; resistance: 0-6K ohms, 60K ohms, 600K ohms, 6M ohms. Volume level: − 22 dB to + 22 dB to + 62 dB in five ranges. Size and weight: 5.5 inches high × 4.3 inches wide × 1.6 inches deep.

Model M-300 is a high quality, highly sensitive, laboratory-type instrument. Its large, easy-to-read scale and excellent damping are usually found only in expensive meters. Parallax errors are eliminated by a mirror arc. This meter comes complete with battery, spare fuse, test leads, and instruction manual. Model M-300, completely wired and tested, $28.95.

Ameco LCD digital multimeter, Model D-200, features high-contrast, large 1/2 inch, 3-1/2 digit LCD display; automatic polarity;



automatic zero adjustment; over-range indication on all ranges; low-battery indication; full overload protection; 10-megohm input impedance; rugged anti-slip case with stand.

Ranges for dc voltage: 0-200 mV, 2V, 20V, 200V, and 1,000V; ac voltage: 0-200V, and 750V; dc current: 0-200 μA, 2 mA, 20 mA, 200 mA, and 10 A; resistance: 0-200 ohms, 2K ohms, 20K ohms, 200K ohms, 2000K ohms, and 20M ohms. Size and weight: 7 inches high × 2.7 inches wide × 1.6 inches deep.

The latest IC and display technology insure reliability, accuracy, and stability. Dual slope integration provides fast, accurate, noise-free measurements. The same two jacks are used for all functions and ranges (except 10A dc). Model D-200 comes complete with battery, spare fuse, test probes, instruction manual, and an optional carrying case. Model D-200, completely wired and tested, $69.95; optional carrying case, $3.75.

For further information, contact Ameco Equipment Company, 275 Hillside Avenue, Williston Park, Long Island, New York 11596. Reader Service Number 305.

RT-1100 receive terminal

DGM Electronics has just introduced the RT-1100 Receive Terminal for Baudot, ASCII, and Morse. The RT-1100 converts the audio from your receiver, decodes it, and displays the words on a video monitor or TV set (using rf modulator). The RT-1100 incorporates an active filter demodulator with scope tuning outputs. It will copy 170, 425, 850 Hz shift RTTY signals at speeds of 60, 66, 75, and 100 WPM on Baudot and 110 baud on ASCII. The unit will copy 6-60 wpm Morse signals using automatic or manual speed tracking.

The RT-1100 has a parallel ASCII printer output for hard copy. The video output provides sixteen lines of thirty-two characters per line with two pages. The second page is stored in memory and can be recalled by using the page 1-2 switch on the front panel. The unit has a built-in 110 Vac power supply and is housed in an attractive 3 × 10 × 10-inch case with brushed, anodized front and rear panels. The cover is a grey wrinkle finish. The unit comes with a one-year warranty on parts and labor.

For more information, contact DGM Electronics, Inc., 787 Briar Lane, Beloit, Wisconsin 53511. Reader Service Number 306.

encoder with ultra thin keyboard

Midian Electronics, Inc., has introduced the TTE-1 TouchTone™ encoder with ultra-thin keyboard. The unit features the thinnest available keyboard/DTMF encoder assembly with automatic PTT and side tone. The keyboard mounts virtually flush on a flat surface. DTMF encoder on the back of the keyboard fits into a 1 × 1-1/2 inch hole for flush mounting. It produces digitally synthesized tones for accuracy and stability with adjustable audio output level and generates twelve standard Bell System TouchTones. Options include keyboard only, without encoder, and LED indicating when automatic PTT is activated.

For more information, contact Midian Electronics, Inc., 5907 East Pima Street, Tucson, Arizona 85712. Reader Service Number 307.

interchangeable antennas

Antenna Incorporated has recently introduced a complete line of interchangeable antennas for use on hand held transceivers and scanners. The Portasuader antennas let the user replace only the radiator section of the antenna while continually reusing the mounting-adapter fitting for the transceiver. The radiators are all internal threaded (No. 10-32) to accept the male thread of the interchangeable mounting adapter. The outer portion of the

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NEW products

socket is etched with the frequency range for that particular radiator. Ten mounts are available to be used with the five different radiator styles.

The short 8-inch whips for 25-54 MHz incorporate a wire-wound base-loading coil and helical-style radiator in six different frequency steps. Tuning has been eliminated and fragile temperature-sensitive ferrite cores have also been eliminated. A distinct feature is the antenna length which is less than 8-inches long, measured from mount to tip. In the 118-174 MHz frequencies, Portasuaders are available in standard tuned helical units, extra-fat helical units and 1/4-wavelength stainless steel whips incorporating a spring section. The advantage of the fat Portasuader is its shorter length (about 2 inches shorter than helical). A secondary benefit in using the fat antenna is its lower Q, broadening the resonance curve and thus achieving a better match over the frequency range.

Also available is a 1/4-wavelength 0.046 inch diameter 17-7PH stainless steel whip incorporating a novel spring construction above the base fitting. This spring allows the whip assembly to bend when the user sits down with his radio attached to his belt. The 1/4-wavelength Portasuader was designed to replace the telescopic antennas, which bend or break or simply do not telescope properly. As a further advantage, the 1/4-wavelength Portasuader antennas exhibit a practical 10 dB gain over the helical or fat helical antennas.

The frequency range is covered in seven frequency steps, thus again removing the need for field tuning. UHF stubby helical whips and 1/4-wavelength speedometer cable antennas are available as radiators in five frequency ranges between 406-512 MHz. An 800 MHz 1/4-wavelength speedometer cable antenna is currently available.

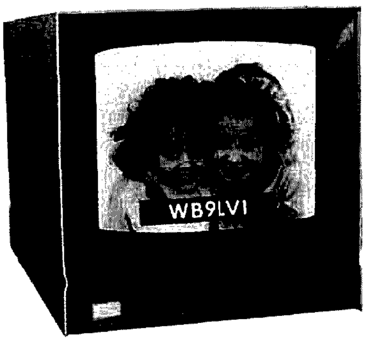
These Portasuaders are constructed from heavy copper-plated spring steel that is screwed onto the base fitting and then soldered to ensure electrical contact. Both helical and speedometer styles are insulated by coating in a multi-stage process. The special process guarantees a solid section of material with minimum voids and high finish gloss. The coating is designed to remain flexible, retain its resilience at −40 degrees F and not to soften at 200 degrees F.

For further information, contact Randy Friedberg, Vice President, Antenna Incorporated, 26301 Richmond Road, Cleveland, Ohio 44146. Reader Service Number 308.

high resolution SSTV converter

High resolution slow scan television (SSTV) is available with the Videoscan 1000 by Microcraft Corporation. The unit is completely compatible with Amateur-standard SSTV and first-generation equipment. Videoscan can convey high-resolution eight-second, 128-line SSTV pictures to first generation scan converters using current standards. In two separate high resolution modes, the TV picture uses the full 256 TV lines and 256 picture elements (pixels) per line, resulting in pictures that rival commercial TV quality. The pixels are quantized to 64 levels of gray, four times better than first generation units. No contouring (false edges) is introduced to detract from the picture.

Some features of Videoscan are: Split-mode, a special mode that enables viewing four regular 8.5-second SSTV pictures at one



time on the TV monitor as they are received; Stop motion, a single frame of video may be grabbed into memory from a TV camera manually or automatically, thus stopping motion; Cursor, a cursor dot appears on the screen to indicate the current line being transmitted; Gray scale, Call Sign, mode selector activates a gray scale and optional call sign which are superimposed on the picture in memory; Station switching, all necessary switching between transmitter, microphone, and tape recorder is included in Videoscan.

Microcraft is presently working on a computer input/output port and a color conversion of the Videoscan 1000.

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NEW products

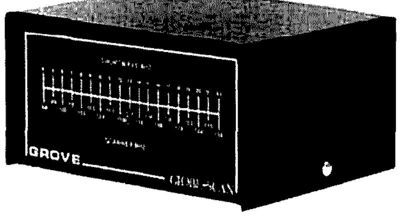
The Videoscan 1000 is available as a complete kit for $595.00 or wired and tested for $795.00 plus $6.00 for shipping.

For more information, contact Microcraft Corporation, P.O. Box 513, Thiensville, Wisconsin 53092. Reader Service Number 309.

improved 225-400 MHz scanner converter

The CVR-1B Scanverter includes a built-in preamplifier for increased sensitivity. It allows complete coverage of the 225-400 MHz military/federal government aircraft band when used with a standard aircraft band scanner. "Bandstacking" allows the entire 175-MHz-wide UHF aircraft band to be compressed into the 118-136 MHz range tunable on any scanner capable of standard aircraft reception. No tuning or adjustments are necessary with the fully automatic converter.

Reception for hundreds of miles is possible when an outside antenna is used. Additional features include high sensitivity, low noise microstripline circuit; all-metal cabinet for superior shielding; double-balanced mixer to reduce images; nine-pole filter to suppress out-of-band interference; crystal oscillator to provide high stability; and Zener diode voltage regulation to limit drift. The scanner is powered by convenient 12 Vdc.

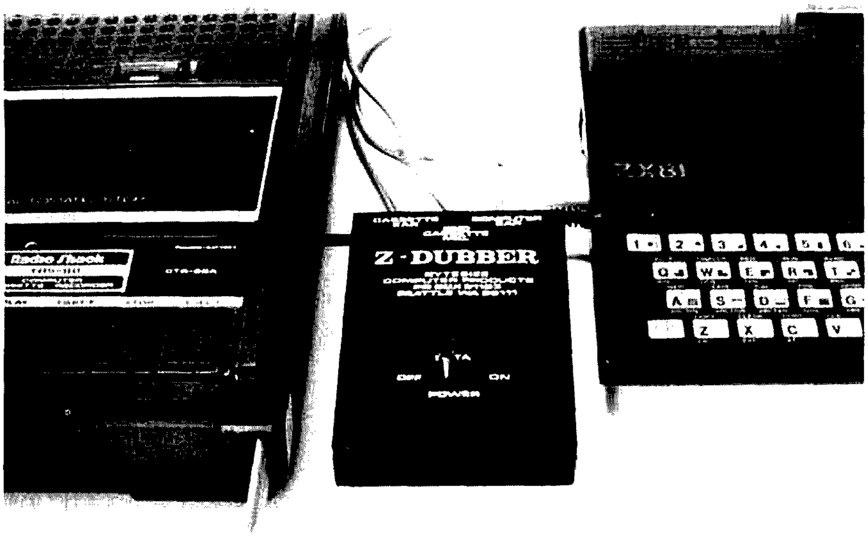


The Scanverter, CVR-1B, costs $89.00 plus $2.00 for shipping. Contact Grove Enterprises, 140 Dog Branch Road, Brasstown, North Carolina 28902. Reader Service Number 310.

the Z-Dubber

The Sinclair ZX81/Timex 1000 is a popular personal computer. One drawback is the difficulty experienced in loading cassette programs. Bytesize Computer Products has introduced the Z-Dubber, an interface between the Sinclair computer and its cassette recorder, which helps even the most difficult cassette program to load easily. Additionally, the Z-Dubber allows you to connect two cassette recorders together to create perfect back-up copies of your Sinclair programs. The Z-Dubber operates on two AAA cells, and is packaged in an attractive black case. It is available for $29.95 plus 3 percent for shipping.

For more information, contact Bytesize Computer Products, P.O. Box 21123, Seattle, Washington 98111. Reader Service Number 311.



Amateur data display, DTU-12

Get a clean, crisp computer-quality data display for your next ham project with a DTU-12 from Dotronix, Inc., available in kit, chassis, or chassis/ac power versions, either P4 (white) or P31 (green) phosphor. It requires only 12 volts at 1.5 amperes, and standard TTL horizontal and vertical control signals with 2.5 volts video drive. The scan rate is 15,750 Hz. Interface is made through ten-pin edge-card connector.

The kit costs $85.00 (CRT/circuit only); chassis $95.00; ac supply $35.00 (for chassis version).

For additional information, contact Dotronix, Inc., 160 First Street S.E., New Brighton, Minnesota 55112. Reader Service Number 312.

programmable CTCSS encoder

A miniature encoder has been introduced by Ferritronics, Inc., featuring quartz-accurate stability and all thirty-seven EIA tones. Two variations are available: the FT303A, which is programmed by cutting wire loops; and the FT303B, which uses a dipswitch for programming. The encoder measures 0.9 × 1 × 0.4 inch and draws less than 7 mA. Mounting holes and color-coded lead set make installation simple.

For further information, contact Tom Whitney at Ferritronics, Inc., 222 Newkirk Road, Richmond Hill, Ontario L4C 3G7, Canada. Reader Service Number 313.

two-meter mobile transceiver

The TR-7950 and TR-7930 are identical in features except for rf output: 45 watts for the TR-7950, and 25 watts for the TR-7930. Their

MANUALS for most ham gear made 1937/1970. Send $1.00 for 18 page "Manual List", postpaid. HI-MANUALS, Box R802, Council Bluffs, Iowa 51502.

SATELLITE TELEVISION — Howard/Coleman boards to build your own receiver. *For more information write:* Robert Coleman, Rt 3, Box 58-AHR, Travelers Rest, SC 29690.

WANTED: Schematics-Rider, Sams or other early publications. Scaramella, P.O. Box 1, Woonsocket, RI 02895-0001.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

RUBBER STAMPS: 3 lines $3.25 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

WANTED: New or used MS and coaxial connectors, synchros, tubes, components, military surplus equipment. Bill Williams, PO #7057, Norfolk, VA 23509.

VERY In-ter-est-ing! Next 5 issues $2. Ham Trader "Yellow Sheets", POB356, Wheaton, IL 60189.

CB TO 10 METER PROFESSIONALS: Your rig or buy ours — AM/FM/SSB/CW. Certified Communications, 4138 So. Ferris, Fremont, Michigan 49412; (616) 924-4561.

HAMS FOR CHRIST — Reach other Hams with a Gospel Tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91788.

"WEST COAST 160 METER BULLETIN" devoted to, for and by top band operators. Edited and published 6 times a year by N7CKD. Subscriptions $7.00 U.S. and U.S. Possessions. Canada and Mexico $7.00 U.S. Overseas rate $8.50 U.S. dollars. To 4248 "A" St. S.E., Box 609, Auburn, WA 98002.

WANTED: Diagram for TPL VHF-FM 120W, 2 meter Amp. — 1002-S. It uses TRW-PT8780 which I have been unable to find. Any help would be appreciated. WA0OFO.

WANTED: Micor and Mstr II Base Stations 406-420 and 450-470 MHz. Also 2 and 6 GHz solid state microwave equipment. AK7B, 4 Ajax Place, Berkeley, CA 94708.

FOR SALE: Ten-Tec Argonaut 515, mint, $280.00; Kenwood TR-7800, exc., $235.00; Icom R-70 receiver, mint, $630.00; R-392, exc., extras, $145.00; add. UPS. Ham mags: SASE for list, details. WA7ZYQ. (208) 245-2070.

PLANS, CIRCUIT BOARDS, AND KIT PARTS (author approved) for Leach's construction projects. Power amps, preamps, pre-preamps and loudspeakers. Send SASE for information. Custom Components, Box 33193, Decatur, GA 30033.

DXPREDICTOR: Computes MUF, FOT, LUF between any two QTHs. Documented; Easy to Use; Nice Graphical/Tabular output. Adapted from Algorithms currently used by USG agencies. Available for: Apple II +/IIe (48kmin) plus DOS 3.3 or cassette); PET/CBM (16kmin, 8050 disk or cassette); VIC20 (16k RAM Card, cassette); CBM 64 (cassette only). Program + Documentation: $40.00 all except VIC20 ($30.00). Documentation only: $5.00 (refundable upon purchase of software). Checks to K.J. Flynn, PO Box 903, Mountain View, CA 94042. (CA residents add 6.5% tax).

SLEP SPECIALS, HP608F late model signal generator, 10 MHz thru 455 MHz, 19" panel with blue/gray cabinet, excellent for precision laboratory work $375.00, URM-25 signal generator 10 kHz thru 50 MHz $265.00, URM-26 signal generator 4 MHz thru 405 MHz $245.00, USM-140 oscilloscope DC-25 MHz, dual trace, triggered sweep $295.00, GRM-46 test set for ARC-27, ARC-55, $40.00, Tektronix 3S76 dual trace plug-in $75.00, Tektronix 1L30 spectrum analyzer plug-in 925 MHz thru 10.5 GHz $495.00, military SG-66/ARM-5 aircraft VOR signal generator equivalent to ARC H-14, perfect for aircraft radio repair $295.00. Add shipping. We accept M/C, Visa or check. Phone 704-524-7519, Slep Electronics Company, Highway 441, Otto, N. C. 28763.

Coming Events

ACTIVITIES

"Places to go..."

CALIFORNIA: The 34th annual International DX Convention, a joint effort of the Northern California and Southern California DX Clubs, Friday, Saturday and Sunday, April 22, 23 and 24, at the Visalia Holiday Inn Hotel, Visalia. DXpedition reports, technical presentations, awards, contests, dining and hospitality rooms, slides, movies

and videotapes, open forums. Many overseas visitors expected. Manufacturers and distributors showing the latest in radio gear. For further information: Northern California DX Club, PO Box 608, Menlo Park, CA 94025.

COLORADO: The Grand Mesa Repeater Society's fourth annual Western Slope Swapfest, Saturday, April 2, 10 AM to 4 PM, Plumbers and Steamfitters Union Hall, 2384 Highway 6 and 50, Grand Junction. Free admission. Swap tables $5.00 each. Auction and refreshments. Talk in on 146.22/.82. For information SASE to Bill Brown, K0UK, 582 So. Maple St., Fruita, CO 81521 or call (303) 858-9661.

GEORGIA: Kennehoochee Hamfest, Sunday, April 17, 8 AM to 4 PM, Civic Center, Marietta, GA.

ILLINOIS: The 17th annual Rock River ARC Hamfest, Sunday, April 10, Lee County 4-H Center, one mile east of jct. 52 and 30, south of Dixon. Doors open 6:30 for dealers; 7:30 general public. 6 ft. tables available $5.00. Advance ticket donation $2, at gate $2.50. Food. Camping available at nominal charge. Talk in on 37/97 repeater. For information and advance tickets: Ed Webb, WD9CJB, 618 Orchard St., Dixon, IL 61021. (815) 284-3811.

LOUISIANA: The Baton Rouge Amateur Radio Club's annual Hamfest, Saturday, May 7 and Sunday, May 8, Catholic High School, 855 Hearthstone Drive, Baton Rouge. Swap tables, dealers, tech forums and activities for non-hams and children. Talk in on 19/79 and 52 simplex. For further information: BRARC, PO Box 4004, Baton Rouge, LA 70821.

MASSACHUSETTS: The Framingham Amateur Radio Association's 8th annual Spring Flea Market, Sunday, April 10; the largest indoor Ham Flea Market in New England. Framingham Civic League Building, 214 Concord St. (Route 126) in downtown Framingham. Doors open at 10 AM, sellers setup starting at 8:30. Admission $2. Tables $10 (pre-registration required). Talk in on 75/15 and 52 direct. Radio equipment, computer gear, bargains galore. For information, tables: Ron Egalka, K1YHM, 3 Driscoll Drive, Framingham, MA 01701.

MASSACHUSETTS: The Wellesley Amateur Radio Society's annual auction, Saturday, April 16, First Congregational Church of Wellesley Hills, 207 Washington Street, Wellesley Hills, intersection of Routes 9 and 16. Doors open 9 AM; auction starts 10 AM. (15% commission, $1.00 minimum, $30.00 maximum). Talk in on 04:64; 63:03; and 52. Contact: Kevin P. Kelly, WA1YHV, 7 Lawnwood Place, Charlestown, MA 02129.

MICHIGAN: S.E.M.A.R.A., The Southeastern Michigan Amateur Radio Association's 25th annual Hamfest Swap and Shop, April 10, 8 AM to 3 PM, Grosse Point North High School, Vernier Road between Mack and Lakeshore. Admission $1.00 advance; $2.00 at door. Good food, free parking. Talk in on 147.75/.15. For information: SEMARA Swap, PO Box 646, St. Clair Shores, MI 48083 or phone Ray Ninness, WD8KXN (313) 777-0119.

MINNESOTA: The Arrowhead Radio Amateur Club's annual swapfest, Saturday, May 7, 10 AM to 3 PM, Holiday Inn, 207 West Superior St., downtown Duluth. Admission $2.50 advance, $3.00 door. Reserved 4 ft. tables $3.50 advance, $4.00 at door. Food, free parking, enclosed shopping mall. Talk in on 34/94. For information, reservations SASE to Jerry Frederick, N0BNG, 1127 - 104th Avenue West, Duluth, MN 55808.

NEBRASKA: The 1983 Midwest ARRL Convention, April 15, 16 and 17, Marina Inn, South Sioux City. Seminars, displays, exhibits and large flea market all indoors. Fine entertainment during Saturday night banquet. QCWA breakfast, 3900 Club luncheon and an outstanding ladies' program Saturday. Convention costs $6.00 for 3 days. Saturday night banquet $10.00 advance; $12.00 at door. To reserve flea market table contact Al Smith, W0PEX, 3529 Douglas St., Sioux City, IA 51104. Exhibitors contact Jim Boise, KA0GZY, 22 LaSalle St., Sioux City, IA 51104. For general information contact Dick Pitner, W0FZO, General Chairman, 2931 Pierce St., Sioux City, IA 51104. For advance banquet tickets and motel reservations contact Jerry Smith, W0DUN, Akron, IA 51001.

NEW ENGLAND: The Hosstraders will hold their tenth annual Tailgate Swapfest, Saturday, May 7, sunrise to sunset, at Deerfield, NH, Fairgrounds. Admission $1.00, including tailgaters and commercial. Friday night camping for self-contained rigs at nominal fee. None admitted before 4 PM Friday. Profits benefit Boston Burns Unit of Shriners' Hospital. Last year's donation $2622.75. Questions or map to northeast's biggest ham flea market? SASE to Norm, WA1IVB, RFD Box 57, West Baldwin, ME 04091 or Joe, K1RQG, Star Route, Box 56, Bucksport, ME 04416 or Bob, W1GWU, North Walton Road, Seabrook, NH.

NEW JERSEY: The 8th Trenton Computer Festival, Saturday and Sunday, April 16 and 17, 10 AM to 5 PM, Trenton State College, Trenton. Exhibits, electronics flea market, technical sessions, free short courses on Sunday. Admission $5. ($3 students). For further information: TCF-83, Trenton State College, Hillwood Lakes CN550, Trenton, NJ 08625. (609) 771-2487.

NEW JERSEY: Annual Flemington Hamfest, Saturday, April 9 from 8 AM to 4 PM at the Hunterdon Central High School Field House. 20,000 square feet of heated indoor area. Gigantic flea market, 200 tables, major manufacturers and more. Bring the family and your friends. Flemington is located between NYC and Philadelphia at the intersection of routes 202 and 31 just 10 miles south of I-78, and is a tourist area. Talk-In 146.52, 147.375, 147.015, 224.12 and 224.54 MHz. Admission $3.00 donation. For reservations or information call (201) 788-4080 or write: Cherryville Repeater Association c/o W2FCW, Box 76, Fairview, Dr., Annandale, NJ 08801.

NEW YORK: The Suffolk County Radio Club's indoor Flea Market, Sunday, May 1, 8 AM to 3 PM, Republic Lodge No. 1987, 585 Broadhollow Road, Melville, Long Island. Admission $2.00 (spouses and children under 12 free). Sellers tables $7.00, includes one admission. Free parking. Refreshments. Talk in on 144.61/145.21 and 146.52. For information: Richard Tygar, AC2P, (516) 643-5956 evenings.

NEW YORK: The 24th annual Southern Tier Amateur Radio Club's Hamfest, Saturday, May 7, the Treadway Inn, Oswego. Flea market opens at 8 AM. Vendor displays and sales; tech and non-tech talks; refreshments. Advance tickets only for a dinner at 6:30 PM. Talk in on 22/82, 16/76 or 146.52 simplex. For further info SASE to KF2X, C. England, RD #1, Box 144, Vestal, NY 13850.

NORTH CAROLINA: The Raleigh Amateur Radio Society's 11th annual Hamfest, Sunday, April 17, Crabtree Valley Mall, U.S. 70 West, starts 8 AM. Admission $4.00 includes tailgating. Tables available for rent. Covered Flea Market. CW and homebrew contests, special interests meetings. Talk in on 04/64, 28/88. For information: RARS Hamfest, PO Box 17124, Raleigh, NC 27619.

OHIO: The 14th annual B*A*S*H Friday night of Dayton Hamvention, April 29, Convention Center, Main and Fifth Streets. Adjacent parking. Free admission. Refreshments and entertainment. Two exciting top awards and more. For further information contact the Miami Valley FM Association, PO Box 263, Dayton, Ohio 45401.

OHIO: The Athens County ARA annual Hamfest, Sunday, May 15, Athens City Recreation Center, U.S. 33 and 50. 8 AM to 4 PM. Setup 7 AM. Outdoor paved flea market space $2.00. Some indoor space available $3.00. Food, free parking, restaurants and recreation area within walking distance. Athens Mall next door. Tickets $1 advance, $2 at gate. Talk in 146.34/.94. Tickets and info: ACARA, PO Box 72, Athens, Ohio 45701. Or call Joe, WB8DOD (614) 797-4874.

PENNSYLVANIA: The first annual Southern Alleghenies Hamfest, April 10, 8 AM to 5 PM, Bedford County Fairgrounds, intersection of Routes 30 and 220. Sponsored by the Bedford ARC; Altoona (Horseshoe) ARC; Cumberland, MD, ARC; Somerset ARC and the Blue Knob Repeater Association, this Hamfest features computer demonstrations, displays, ARRL booth, refreshments and more, all in a large heated building. Talk in on the Bedford repeater 145.49 and 146.52 simplex. Admission $3.00. Inside tables $5.00. For information: Tom Gutshall, W3BZN (814) 942-7334 or on the 147.75/15 Blue Knob Repeater.

ROCHESTER HAMFEST: Atlantic Division/New York State Convention. Saturday, May 21, Monroe County Fairgrounds. Hotel headquarters, Rochester Marriott Thruway. More info? Write or call Rochester Hamfest, 300 White Spruce Blvd., Rochester, NY 14623 (716) 424-7184.

SOUTH CAROLINA: The Blue Ridge Amateur Radio Society's Hamfest, Saturday, April 30 and Sunday, May 1, at the American Legion Fairgrounds, White Horse Road, Greenville. Admission $3.00. Talk in on 146.01/61 and 223.46/224.06. For information: Phil Mullins, WD4KTG, Hamfest Chairman, PO Box 99, Simpsonville, SC 29681. For advance sales: Mrs. Sue Chism, Rt. 6, 203 Lanewood Dr., Greenville, SC 29607.

TEXAS: TARS, The Tidelands Amateur Radio Society's Springfest 1983 at the fairgrounds in League City, Saturday, April 16. Auction, displays, demonstrations, good food and fellowship. Free admission. Refreshments available from 7 AM. Activities start 9 AM to 4 PM. For information: T.A.R.S., PO Box 73, Texas City, TX 77590.

WASHINGTON: The Central Washington State Hamfest sponsored by the Yakima Amateur Radio Club, W7AO, Saturday, May 14, 9 AM to 5 PM, lunch available, and Sunday, May 15, 8 AM to 2 PM, breakfast and lunch, the Hobby Building, Central Washington State Fairgrounds, Yakima. Combination ticket $4.00 advance; $5.00 door. Additional tickets 2/$5.00. Regional dealer displays and FREE swap and shop with plenty of tables. Talk in on 146.01/61. For tickets and information: Dan Haughton, PO Box 9211, Yakima, WA 98909.

WISCONSIN: The Madison Area Repeater Association's 11th annual Swapfest, Sunday, April 10, Dane County Exposition Center Forum Building, Madison. Doors open 8 AM for commercial exhibitors and flea market sellers; 9 AM for general public. Admission $2.50 advance and $3.00 door. Children twelve and under free. Flea market tables $4.00 each advance and $5.00 door. Reserve early. Exhibitors and flea market sellers will have equipment and components for hams, computer hobbyists and experimenters. An all-you-can-eat pancake breakfast and a Bar-Bar-Q lunch will be available. Talk in on WR9ABT, 146.16/.76. For information and reservations write M.A.R.A., PO Box 3403, Madison, WI 53704.

WISCONSIN: The 3F ARC Swapfest, May 7, 8 AM to 3 PM, Neenah Labor Temple. 4 ft. tables $1.50 advance; $2.00 at door. Talk in on 144.61/145.21. For advance registration: Mark Michel, W9OP, 339 Naymut St., Menasha, WI 54952.

WISCONSIN: The Ozaukee Radio Club's 5th annual Swapfest, Saturday, May 7, 8 AM to 1 PM, Circle B Recreation Center, Highway 60, Cedarburg, 20 miles north of Milwaukee. Admission $2.00 advance, $3.00 door. 8 ft. tables $3.00 each. Food and refreshments. Sellers admitted at 7 AM for setups. For tickets, tables, maps or information SASE to Ozaukee Radio Club Swapfest, PO Box 13, Port Washington, WI 53074.

OPERATING EVENTS

"Things to do..."

APRIL 6, 7 AND 13, 14: DX-YL to North American YL Contest. CW: Wednesday, April 6, 1800 UTC to Thursday, April 7, 1800 UTC. Phone: Wednesday, April 13, 1800 UTC to Thursday, April 14, 1800 UTC. All licensed women operators throughout the world are invited to participate. DX YL call "CQ North American YL". N.A. YL call "CQ DX YL". All bands may be used. No cross band operation. Net contacts, repeater contacts and contacts with OMs do not count. Stations may be worked/counted once on each band and mode. Exchange: Station worked, QSO number, RS(T), state or country. Entries in log must show time, band, date and xmitter power. Phone and CW scored as separate contests. Submit separate logs for each contest. DX-YLs incl. Hawaii and Alaska, may contact all N.A. continent which includes 48 cont. states and Canadian Provinces. Contestants on N.A. continent may contact DX stations to include Hawaii and Alaska. A station may be counted once on each band for credit and one point is earned for each station worked once on each band. Multiply number of QSO's by number of different states and provinces or countries worked. A multiplier is counted once in contest, NOT on each band. Contestants running 150 watts or less on CW and 300 watts PEP or less on SSB, may multiply results by (E) by 1.25 (low power multiplier). Logs must be signed by operator and postmarked by April 28, 1983 and received NLT May 23, 1983. Send logs to YLRL Vice President.

APRIL 17 AND 18: The Central Massachusetts Amateur Radio Association will commemorate Patriot's Day, honoring the Minutemen and other patriots who fought during the American Revolution. Club station W1BIM will operate Sunday, 1700 UTC to 2200 UTC; and Monday, 1500 UTC to 2200 UTC from the Worcester, Mass. Science Center, approx. 20 kHz up from the General phone band edge, 40-10 meters. For a special certificate send QSL card and business SASE to: Alan Freeman, KA1XL, 83 Newton Avenue North, Worcester, MA 01602.

APRIL 22-24: A special events station, K0TIK, will operate from the Nebraska State Arbor Lodge, former home of Arbor Day founder, J. Sterling Morton, in Nebraska City, NE. Tree City U.S.A. during the annual Arbor Day celebration. This station plus other club member stations will operate in the general portion of phone and CW bands, 80-10 meters, 2400 UTC Friday to 0600 UTC Sunday. All Amateurs contacting this or any other club member station will be eligible to receive an Arbor Day commemorative certificate. Send one dollar and business SASE to N.C. A.R.C., Box 8, Nebraska City, NE 68410.

APRIL 23 AND 24: TSRAC Scavenger Hunt Contest, 0000Z, April 23 to 2359Z, April 24, sponsored by the Triple States Radio Amateur Club. Two trophies to be awarded; one to General Class and above licensee with highest score and one to Novice or Technician Class operator with highest score. Second and third place certificates awarded also. Modes: CW and phone. Exchange: QSOs or "CQ TSHT TEST", 20 kHz ± above bottom of any General or Novice band. Open to all Amateurs. Submit entries to contest chairman: David M. Kinney, KC8YR, RD #1, Mingo Jct., OH 43938 by May 25.

APRIL 23 AND 24: The Independent Amateur Radio Group of Delaware will operate from atop the U.S. Geological Survey marker at the point where Pennsylvania, Maryland and Delaware meet, Mason and Dixon's Stone Number one, from 1500Z to 2300Z each day. Rain date April 30 and May 1. 10 through 40 meters in lower ends of General segments. 2M activity announced on local repeaters. ARRL assures QSL cards will count for any or all three states. Each operator will use own call with a "/3/3/3" identifier. Special QSL card for SASE to operator worked.

APRIL 23 AND 24: QRP Amateur Radio Club International Spring QSO Party, Saturday, 1200 UTC to Sunday, 2400 UTC. Exchange: Members give RS(T), state/prov-

ince/country and QRP ARCI membership number. Non-members give RS(T), state/province/country and power output. QSO points (total all bands) times total number of states/provinces/countries (may be worked on more than one band) times power multiplier times bonus multiplier (if any) equals claimed score. Send large SASE or IRCs to contest chairman for scoring summary sheet in advance of contest. Send full log data plus separate worksheet showing details and time off air. No logs returned. For results and scores send large SASE with one ounce of U.S. postage or IRCs. Logs must be received by May 21, 1983. QRP ARCI Contest Chairman, William Dickerson, WA2JOC, 230 Mill St., Danville, PA 17821.

APRIL 23 AND 24: The Missouri Valley Amateur Radio Club's fourth annual Pony Express Day, 1000 CST to 1900 CST (Saturday) and 0900 CST to 1200 CST (Sunday). This event commemorates the original running of the Pony Express from St. Joseph, Missouri to Sacramento, Calif. Operating frequencies: 10 kHz from bottom of the general phone bands on 15, 20, 40 and 75 meters. On 10 meters — 28.575. CW: 10 meters — 28.150; 15 meters — 21.150; 40 meters — 7.125. Anyone contacting Club station WØNH is eligible for a special Pony Express certificate. Just send two first-class postage stamps and a QSL card to: Missouri Valley Amateur Radio Club, 401 N. 12th Street, St. Joseph, MO 64501.

APRIL 29-MAY 1: The first International VHF/UHF Conference to be held as part of the Dayton Hamvention. Activities span all three days and include tech talks and forums; noise figure and antenna gain measuring contests, a hospitality suite get-together with refreshments. All this along with the rest of the Hamvention features. For further information and to advise of participation in contests contact: Jim Stitt, WA8ONQ, 311 N. Marshall Road, Middletown, OH 45042. (513) 475-4444 business or (513) 863-0820 home.

MAY 7: Harry's Haydays. The Southside Amateur Radio Club will operate KAØHXU to commemorate President Harry Truman's 99th birthday. The station will operate at or near the old Truman farm home in Grandview, MO from 1500Z to 2400Z on 21.355, 14.290 and 7.230. Commemorative QSL's will be sent via the bureau unless otherwise requested. For information: Southside ARC, PO Box 412, Grandview, MO 64030.

WORKSHOP: Personal Computer Interfacing and Scientific Instrument Automation. $395.00. Charlotte, NC, June 2-4; Reston, VA, June 16-18; Charleston, SC, July 14-16; Williamsburg, VA, Aug. 11-13; and Greensboro, NC, Sept. 8-10. These are hands-on workshops with each participant wiring and testing interfaces. For more information, call or write Dr. Linda Leffel, C.E.C., Virginia Tech, Blacksburg, Virginia 24061. (703) 961-4848.

MAY 1983 / $2.50

ham radio magazine

- log-Yagis simplified
- 20-meter linear array
- inexpensive hardline connectors
- 20-meter mobile vertical
- stagger-tuned dipoles

focus on communications technology

ham radio
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contents

REFLECTIONS

Antennas ... Antennas ... Antennas

Our May issue has historically been our antenna issue and this year will be no exception. Within the pages of this, our fattest issue in a long time, are twelve articles on antennas of all sizes, shapes, and applications. Let me, if I may, be your guide through the next 136 pages or so and provide you with a summary of the various articles contained within.

We start with part one in a series of articles on phased vertical antennas by Forrest Gehrke, K2BT. Forrest, you may recall, provided us with "A Precision Noise Bridge" in the March, 1983, issue of *ham radio*. He, like so many of us, is not satisfied with copying previous designs and letting it go at that. He must look into things, carefully examining the technical reasons for the correct operation of devices. It is with that approach that he examined the interrelated properties of phased vertical arrays — perhaps more closely than has ever before been done in the pages of a ham journal.

Part one of the series explores *incorrect* assumptions accepted by many (and unfortunately used by many) in their designs of antenna systems. Foremost among these incorrect assumptions is the concept that mutual coupling between elements can be ignored. Following close behind is the argument that, if an array requires equal current drive, then driving each element with equal *power* will always satisfy that requirement. Forrest leads us from the theoretical design to actual drive-network hardware, and shows us how a repeatable 30 to 40 dB F/B ratio is achieved. Like W2PV's series on Yagis, Forrest's series on phased verticals will be both interesting and useful.

K4MT shows us how it is possible to use one wire array over the widest (percentage bandwidth) Amateur band while not exceeding a 2:1 VSWR. His stagger-tuned dipoles cover the 13.3-percent-wide 80-meter band, producing a W-shaped SWR curve. This same design technique is easily applied to 160, 40, and 10 meters with their percentage bandwidths of 10.5, 4.2, and 5.9 percent, respectively.

For a change of pace, a few shorter articles by K9CZB, AA6PZ, W6SAI, and WA8DXB illustrate interesting ways of providing superior performance with little expenditure of time or money. K9CZB shows how an auto replacement antenna and a CB whip can combine to give broadband, durable mobile capability on the 20-meter band. AA6PZ illustrates three different 2-meter antennas or improvements that are lightweight and easy to build. His last design is a 10-dB-gain collapsible four-element Yagi. This weekend project will help you raise those distant repeaters that your handheld previously struggled to access. W6SAI brings us back to basics with his discussion on the various shapes and gains associated with loop antennas. He provides, in "Ham Radio Techniques," design data for two-element quads for the 10, 15, 20, and 40 meter bands. WA8DXB, in order to increase his station performance to Asia, reproduces a four-element 20-meter collinear that holds its own against some impressive high-gain Yagis — without going above 16 feet.

W7DHD brings us back to verticals with his examination of five different 1/16-wavelength-high shortened verticals. He compares top loading, top and base loading, center loading, and base-only loading. He quantitatively shows us how to compute the relative field strength of each antenna with respect to a reference quarter-wave, without actually erecting any antennas. An eye-opener is his calculation showing a difference of over 20 dB in performance between a base-loaded vertical and its full-sized quarter-wave counterpart.

John Belrose, VE2CV, a name familiar to many of us, walks us through a design of a highly efficient radiator known as a grounded monopole with elevated feed. This off-center-fed antenna is useful on six Amateur bands (for that matter, it can be used over the entire 3-30 MHz hf spectrum) and does not require traps. Its best feature is that it produces low-angle radiation at all frequencies.

W3EB explains the significance of his 10/11/12 number sequence in his article "Log-Yagis Simplified." Imagine a 10-meter antenna that achieves 11 dB (dipole) gain using only a 12-foot boom! He shows how, and creates even more interest in his longer beam designs with up to 15 dBd gain. He emphasizes the importance of maintaining close tolerances and using careful workmanship.

Broadcasters have been doing it for years: K3ED, in borrowing some of the same principles, shows how to produce steerable nulls with theoretically infinite attenuation in his article "Achieving the Perfect VHF Antenna Null." Construction details are provided for a trombone-type, adjustable-length phase line made from readily available hobby shop brass tubing supplies. It, and a variable-amplitude JFET preamp, are the main components for an electronically controllable antenna system of extremely high F/B ratio. If a particular direction must be locked out or nulled (as in some repeater applications), the same electronic-control can be used in a transmitting array.

Also on the subject of repeaters, K7NM shows how a low-signal condition known as shadowing can be reduced by judicious choice of the high site antenna. His rugged four-pole collinear uses progressive phase delay sections to tilt the beam pattern downward. This reduces overshooting the desired coverage area and cuts back on wasted higher-angle radiation from the same array. The article "Repeater Antenna Beam Tilting" is worthwhile reading for all clubs considering new or improved repeater site constructions.

Rounding out this issue is an article by WB4GCS entitled "Inexpensive Connectors for Hardline." With $2.00 worth of plumbing materials and ten minutes of labor you can build extremely low-loss homemade connectors to use with the surplus 1-inch (2.54-cm) CATV hardline cable now becoming available to hams at low cost. VHF and UHF enthusiasts can now use this high-quality, low-loss cable for repeaters or home stations, without the cost of expensive connectors.

Marty Hanft, KA1ZM, the editor of *ham radio*, is taking his leave, after five years with the magazine, to spend some time overseas. He joined the staff as administrative editor in 1978, working closely with the late Jim Fisk, and has continued providing us with his inimitable editing and organizational talents. We wish him all the best in his new endeavors.

Welcome aboard is extended to Dorothy Leeds, our new assistant editor. Dorothy brings with her technical-magazine editorial and production skills that will be constantly called upon for our rapidly growing Amateur technical magazine.*

Keep those letters coming. Our technical forum and correspondence departments are growing as a direct consequence of the interest shown in the past few months. Please be patient with us — the flood of mail has created a little backlog — but we love it.

Rich Rosen, K2RR
Editor-in-Chief

*This issue of *ham radio* is 42 percent larger than last January's issue.

presstop de W9JUV

20 METER U.S. PHONE EXPANSION WAS APPROVED BY THE FCC at its March 31 Agenda Meeting. The new bottom edge will be 14150, with an exclusive Extra slot from 14150 to 14175. The Advanced Class lower edge moves to 14175, while the General portion now starts at 14225.
Expansion Of The Other HF Amateur Bands Was Held Off by the Commissioners, to be considered in a later NPRM. When the new 20-meter frequencies will actually be available for use hadn't been settled at presstime, but should be about the time you read this.
The 10-Year Amateur License Was Also Discussed at that same meeting. A Notice of Proposed Rulemaking was the result, and though it was not yet ready for release at presstime it's expected to be quite straightforward. The comment due date had not been set as we went to press, but the comment period should be short as little controversy is likely to be raised by this proposal.

COMMENT DUE DATE ON THE FCC'S "NO-CODE" LICENSE proposal has been extended to June 28. The FCC agreed to a 60-day extension of the Comment deadline on the request of the ARRL, whose next Board of Directors' meeting is set for late April. The League's final position on that bitterly contested issue won't be set until that meeting, and the original April 29 Comment due date would not have allowed the League enough time after the meeting to prepare and submit its comments.
Organized Opposition To A "No-Code" License has been developing in several areas. "Grassroots" anti-"No-Code" groups are reported active on both coasts, and one is seeking a spot on the Dayton Hamvention program to rally sentiment against the proposed new license.

THE PHASE 3B SATELLITE LAUNCH IS STILL SET FOR MID MAY, and the European Space Agency is still confident that the trouble-plagued Ariane rocket's problems have now been solved.
Don't Expect To Be Able To Use Phase 3B Right Away, even if the launch is on schedule and trouble free. Checkout and stabilization of the new bird could take several weeks or more, before Amateurs will be able to enjoy the benefits of its elliptical orbit.

AMATEUR OUTRAGE OVER N6BHU'S LICENSE REINSTATEMENT after it had been lifted by the FCC for "profane and indecent" language has now reached Congress. Sen. Barry Goldwater, K7UGA, challenged FCC Chairman Fowler about the FCC Review Board decision to return the violator's license at a recent Senate Communications Subcommittee meeting, and was promised the controversial action would be reviewed by the Commissioners at an early date.
N6BHU Has Promised To Take The Fight Into Federal Court if the Commission decides again to suspend his license. In a conversation with Westlink's WA6ITF, N6BHU said he'd go all the way to the Supreme Court if necessary to keep his Amateur license.
ARRL Has Also Formally Intervened In The N6BHU Case, concerned that the Review Board set an "unlawful and intolerable" precedent in its decision that the language N6BHU had used on the air was acceptable in the Amateur service.

ARTHUR GODFREY, K4LIB, PASSED AWAY MARCH 16 in a New York hospital from pneumonia. He was one of the nation's best known Amateurs, having been a top rated broadcast entertainer for many decades. Arthur, who was 79, narrated "The Ham's Wide World" in 1969 and had been co-narrator of "The World of Amateur Radio" in 1979.

STANDARDS FOR RF RADIATION SHOULDN'T BE THE FCC'S PROVINCE, an all-industry group agreed at a meeting with key Commission people March 16, but the FCC will have to fill a void until the Environmental Protection Agency can complete its RF studies and take on the responsibility. That's still probably two years off, and until then the FCC is expected to use the 10 mW/square centimeter 1982 ANSI standard (with reductions at frequencies to which the body is most susceptible) as meeting the requirement of the EPA. A major benefit will be federal preemption of proliferating state and local RF exposure regulations.

K5LFL's 2-METER OPERATION FROM THE SPACE SHUTTLE is now almost certain, following NASA's OK of the proposal. The only approval still required is from the European Space Agency, whose space lab will be the Shuttle's cargo for that late September launch.

A NEW 220-MHZ DX RECORD WAS SET MARCH 9 when KP4EOR worked LU7DJZ, a 3670 mile QSO. KP4EOR used both SSB and CW for the record-breaking trans-equatorial-propagation contact, while LU7DJZ used CW only. The previous 220 MHz record was 2540 miles between W6NLZ and KH6UK, set back in June, 1959.

CABLE TV CHANNEL E WON'T BECOME A PROBLEM TO 2-METER users in some parts of Chicago. Several of the successful bidders for the multi-area Chicago cable TV franchise, including Continental Cablevision, voluntarily agreed as part of their proposals to give up service on either channel E (2 meters) or K (220 MHz).

A $1000 FINE HAS BEEN LEVIED AGAINST A BURBANK (ILLINOIS) Amateur who recently erected a new 34-foot tower. The Amateur, a minister and former missionary in Nigeria, was anxious to resume contact with former colleagues. Court action on Burbank is still hanging fire.

ham radio Book Review

ham radio magazine takes pleasure in providing the following reviews of books pertinent to Amateur Radio.

rf circuit design

The first word that comes to mind in reviewing the book *rf circuit design* by Chris Bowick, WB4UHY, is *practical*. The author has accomplished in this book what many more-expensive volumes have not been able to — he has provided in one 176-page volume a useful collection of material on rf techniques.

Most rf designers will probably agree that their knowledge took years to acquire and sometimes required access to many different volumes to understand even a single concept. Chris gets right into the essential aspects of each subject, using clearly defined terms, charts, and examples. An elementary example of this is seen on the first page of chapter one, Components. A chart on wire sizes shows how one can quickly determine unknown wire diameters if it's remembered that No. 50 AWG is 1 mil and doubles for each six wire sizes. No. 44 AWG has a 2 × 1, or 2 mil, diameter.

This book, useful to hams who are interested in designing their own equipment, provides numerous examples for guidance each step of the way. There are seven chapters, labelled: Components, Resonant Circuits, Filter Design, Impedance Matching, The Transistor at Radio Frequencies, Small Signal RF Amplifier Design, and RF Power Amplifiers.

There are also three additional sections: Appendix A, use of complex numbers, recommended for those who are not familiar with complex number arithmetic; Appendix B, noise calculations, a systems approach to low-noise design; and Appendix C, bibliography of technical papers and books related to rf circuit design. These additional sections complement this already useful book with material that enables the interested reader to continue his research.

Published by Howard W. Sams, this book is available soft cover (8½ × 11) from Ham Radio's Bookstore, Greenville, New Hampshire 03048, for $21.95 plus $1.00 shipping and handling.

directional antenna patterns

To this reviewer's knowledge, there is not another book around like *Directional Antenna Patterns* by Carl E. Smith, president of the Cleveland Institute of Radio Electronics. It provides under one cover a collection of 15,160 directional antenna patterns, and has become the bible for a-m broadcast antenna design engineers. With the current increase in interest in phased vertical arrays the Radio Amateur will find this material pertinent in several ways.

Part one contains the theory behind the determination of the size and shape of directional-antenna patterns, starting with the standard reference antennas (uniform hemispherical radiator, vertical current element, quarter-wave verticals) and developing into the generalized equation for a directional *n*-antenna array.

Part two, entitled "Systemization of Two Tower Patterns," provides 568 patterns available from a two-element array, examined at electrical and phase separation steps of 15 de-

grees and 45 degrees. It is worthwhile pointing out that commonly used 90 degree space/90 degree phase, that is, quarter-wave separated, quarter-wave-phase difference verticals are just one of the 568 cases considered. Amateurs who don't have the space to separate their verticals by a quarter-wave can still obtain a cardioid switchable pattern by choosing a different set of parameters.

Part three, Systemization of Three Tower Patterns, furnishes 14,592 field plots with 45 degree incremented spacings out to one wavelength for both antenna 2 and antenna 3. The guide to all these different patterns is provided by a systemization placement chart illustrated on each page of 64 patterns.

Directional Antenna Patterns is available hard bound (8½ × 11) by Carl E. Smith for $22.00, postpaid. Contact Smith Electronics, Inc., 8200 Snowville Road, Cleveland, Ohio 44141.

radio communications receivers

Radio Communications Receivers by Cornell Drentea is a new 280-page paperback book available from TAB Books, Inc. The book is billed as a comprehensive guide to radio receiver design and technology, and includes the history of radio technology as it has affected receiver design over the years.

Mr. Drentea attacks the subject of radio receivers systematically, introducing each aspect of a receiver, the design theory, and construction. He also presents an explanation and alternative routes for reaching the same result. State-of-the-art technology is traced from its more primitive beginnings, and future design trends are introduced.

The book is a blend of theory and application, and is meant as a reference for the design and construction of receivers. Design considerations for modern receivers are thoroughly covered, and include the use of computers. The book should prove to be a handy tool to have in your library.

Radio Communications Receivers is available from Ham Radio's Bookstore, Greenville, NH 03048 for $13.95 plus $1.00 shipping and handling.

digital PLL frequency synthesizers — theory and design

Dr. Ulrich Rohde, a name familiar to many of us, has borrowed from his years of knowledge and experience with synthesizers and produced under one cover a collection of data on this complex yet increasingly important subject: *Digital PLL Frequency Synthesizers – Theory And Design*. As stated by Dr. Rohde, the objective of the book is "to provide as much practical circuit information as possible while presenting only the necessary mathematical background and formulas."

This is accomplished in six chapters starting with Loop Fundamentals. Here he introduces the basic linear and digital loops with formulation provided for type 1 and type 2 — first through third order loops. As an example, in the discussion of a type 2, third-order loop, the transfer function is defined along with its application to the suppression of fm noise in a VCO.

Chapter 2, Noise and Spurious Response of Loops, considers an extremely pertinent and limiting factor in any system that uses a synthesizer — sideband noise. The noise sources indicated are leakage from the reference device in phase-locked loops, incomplete suppression of the unwanted component of the mixer output, and inherent noise from the oscillator.

Chapter 3 deals with special loops, that are basically one-loop synthesizers. Techniques are discussed that simultaneously solve the two major requirements of loop operation: resolution and speed. This leads us to a more sophisticated development

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known as the fractional N phase-locked loop.

The Radio Amateur or experimenter will probably find Chapter 4 most useful. Here the loop components consisting of oscillator, reference standard, mixer, phase/frequency comparator, wideband amplifiers, programmable dividers and loop filters are clearly defined and designs provided. Numerous actual circuits are illustrated complete with component values.

With the first four chapters providing a comprehensive understanding of loops, chapter five introduces the multiloop synthesizer that uses a combination of fractional division N synthesizer, sequential phase shifter and digital frequency synthesis techniques. The Rohde and Schwarz EK070 10 kHz to 29.99 MHz shortwave receiver incorporates several multiloop synthesizers and provides a working example of this modern loop concept.

Chapter six finishes this discussion on digital PLL frequency synthesizers with three practical circuits: a) A single-loop, 1-kHz reference synthesizer operating from 41 to 71 MHz, used in a simple shortwave receiver; b) A fast, single-loop 25-kHz synthesizer operating from 41 to 71 MHz; and c) A low sideband noise multi-loop synthesizer covering 75 to 105 MHz in 100-Hz increments.

The appendix includes a mathematical review including a very useful table relating real-time functions with their LaPlace transforms. As indicated at the beginning of this review, Dr. Rohde has generated a 494-page compilation on the current state-of-the-art in digital PLL frequency synthesizers that is useful to the engineer or anyone else who needs a detailed working knowledge of these techniques.

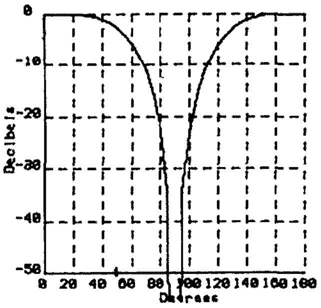
This book, published by Prentice-Hall, is available in hard cover from Ham Radio's Bookstore, Greenville, NH 03048, for $60.00 plus $2.00 shipping and handling.

ham radio

short circuits

Bobtail curtain

The March, 1983, article by Woody Smith, "Bobtail Curtain Follow-Up," indicates that the half-power beamwidth of the Bobtail is 50 degrees. However, using a Sharp PC1500 and assuming a 1:2:1 current ratio, that calculates out to approximately *100* degrees. Note that only *half* the azimuth plot is shown in **fig. 1**; this symmetrical field pattern has a mirror image to its left.



Ed. note: Woody says he took only half of the 3-dB beamwidth numbers for both the Bobtail and half-square. The half-power beamwidths are 100 degrees and 120 degrees, respectively.

vertical phased arrays:
part one

Rotatable arrays for the low bands

Forrest Gehrke, K2BT, with two hundred and fifty-two countries worked on 75 meters, has, over the years, followed a natural progression from the use of simple antennas on the low bands to his present 4-square array. This first installment in a multipart series will help dispel some of the myths associated with phased array design. Though many of the statements might at first glance appear obvious, I *cannot* stress enough the importance of carefully reading this introduction. As Forrest aptly states, a phased array design is not black magic. Achieving outstanding performance just requires a clear understanding of the mechanisms involved. Ed.

Many DXers get on the low bands, if they do at all, to fulfill an award requirement. A low inverted-V or dipole is pitched up, the necessary QSLs collected, and then it's back to the HF bands. But some get hooked and stay. They relearn what the radio pioneers discovered: The low bands are a highly predictable and reliable means of long-distance communications, and, in low sunspot periods such as we are now entering, they're the only after-sunset DX game in town. Sorely missing is directional ability, such as even a modest tribander can provide in the HF bands.

Even if it were practical to rotate that low inverted-V or dipole, it would remain a sad fact that most of the signal is radiated at very high angles with virtually no azimuthal directivity. The result is that the impression easily might be gained that the low bands are good for 500 to 1000 mile contacts but no real DX — that is, until the newcomer happens to eavesdrop on one side of a real DX contact. Then he is amazed to hear a Q5 report given, and at the turnover hear nothing except noise. The old adage "You can't work 'em if you can't hear 'em" is particularly apt on the low bands, where atmospheric static as well as manmade noise is very high.

restricting noise pickup

How is it possible to get a low radiation angle and still beat the noise problem? Perhaps this question seems a contradiction because, as the radiation angle is lowered, the paths over which the antenna receives major noise sources are lengthened, whether the noise is manmade or natural. We may not be able to restrict noise pickup in the paths of interest, but we can at least reduce it from undesired paths with a directional array. On the low bands atmospheric noise is very often quite markedly directional, and it is not unusual to find noise levels differing by 30 dB or more between *various quadrants* of the horizon. Experience shows that high F/B ratio, that is, superior rejection of signals from undesired directions, has far more importance than gain on the low bands for this reason.

It is well known that for reliable DX work a horizontally polarized antenna array had best be one-half to

By Forrest Gehrke, K2BT, 75 Crestview, Mountain Lakes, New Jersey 07046

two wavelengths above the ground for optimum radiation angle. At 20 meters and shorter this is not too difficult, nor is rotating the antenna, but for 80 or 160 meters such heights become impractical — and rotation is virtually impossible.

One obvious alternative is a vertical antenna with electronic directional control. If such an antenna is combined with a good ground plane, one can get radiation angles as low as those possible with a horizontal antenna two wavelengths above ground. But doesn't a vertical "radiate equally poorly in all directions"? And isn't it said to be noisy? After all, everyone knows that, for some mysterious reason, man-made noise sources are supposed to radiate with vertical polarization. That a vertical's very low radiation angle may have something to do with this is seldom considered.

Widespread misinformation on the vertical antenna in Amateur publications is a serious problem. Recently I researched respected Amateur publications printed since 1970, looking for articles on the vertical that contained definitive technical data. I found only two, one quoting the typical dissimilar and reactive driving impedances of the elements of a two-vertical array,[1] and the other calling attention to the need for maintaining unity current ratio despite this dissimilarity.[2] No quantitative data was available for arrays with more than two elements. A few writers included qualitative comments on the vertical array, indicating awareness of the complexity of the matching situation, but most did not. Perhaps this is because, unlike many horizontal arrays, vertical arrays are often designed with all elements driven, thus making the job of satisfying drive current and phase conditions more complicated.

mutual coupling

At this point it may be useful to review the gain mechanism of a Yagi.[3] The Yagi creates gain in the favored direction as a result of the driving currents and phase currents induced in the parasitic elements by means of mutual coupling between the driven and parasitic elements. With appropriate spacings and lengths chosen for the design frequency, current and phase are caused to exist in each element such that the signal is reinforced in the forward direction and partially cancelled in the other directions. The single driven element will present a significantly lower impedance than it would as a lone dipole, because of the loads coupled to it from the parasitic elements. If a low VSWR is not a goal, this element may be driven directly without affecting the gain pattern of the array. The presence or lack of an impedance transformer (such as a Gamma match) has nothing to do with the gain pattern — only with the match to the feedline. A comparison of the current and phase at the midpoint of each element with respect to the driven element, shows that the current magnitude ratio is below unity (about 0.2 to 0.5.), generally rising or falling in each succeeding parasitic element. The phase angle will lead in the reflector (because this element is longer than a half-wavelength); it will lag at the directors (because they are shorter than a half-wavelength), the angle lagging more in each director as we move toward the front of the array. The interaction is quite complex, since there is mutual coupling among the parasitic elements as well as with the driven element. Nevertheless, it is this phenomenon of mutual coupling that permits us to produce directionality in multi-element arrays.

While it's true that driving each element provides an additional controllable variable, this does not mean that no other drive source is acting on the elements. The same mutual coupling that occurs in the Yagi is present here and must be taken into account as part of the total drive to each element. To illustrate, suppose you want to drive an element of an array with 1 ampere at 90 degrees lagging angle. Assume that, at the same termination impedance of this element, mutual coupling from other elements is inducing 0.8 ampere at 90 degrees lagging. An additional drive current of only 0.2 ampere at 90 degrees lag would be all that's needed. In practice, of course, mutual coupling and this additional drive from the feed network may not add arithmetically. Phase angles probably will be different, resulting in vectorial addition. There's another real life complication: The added drive changes the mutually coupled drive! In fact, changing anything at all changes all the other variables because the mutually coupled elements and feed network are all part of one coupled system. This is why the element driven impedances are referred to as driving-point impedances; they exist only while connected to the feed network. We cannot disconnect any element and verify its value with an impedance bridge.

The assumption that mutual coupling doesn't occur (or isn't important) is a mistake found in many articles on phased arrays, vertical or horizontal, in the Amateur publications. This error is almost invariably compounded by a second and more erroneous one: Electrical length of the delay line is equated to current delay in all circumstances, (for example, a quarter-wavelength line is assumed to produce a 90-degree delay regardless of its termination). But equating electrical length to current delay holds true only under certain conditions:*

1. For any length if terminated by a *pure* resistance equal to the characteristic impedance of the line.

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*Except when specifically noted, only the lossless cases will be considered. At low-band frequencies, losses normally are negligible. Calculations including them add greatly to complexity while resulting in insignificant benefit.

2. For an odd number of quarter-wavelengths if terminated by a *pure* resistance of any value.

3. For any number of half-wavelengths regardless of termination impedance.

4. In some special cases (normally of no concern in these applications).†

Disregarding mutual coupling leads to inaccurate results, particularly as regards front-to-back ratio. The designer who makes this error is also typically led to some or all of the following subsidiary assumptions:

1. That the driven impedances of each element always are equal.

2. That if the elements are resonant, the driven impedance of each element is resistive.

3. That if array feedlines are quarter-wavelength, a 90 degree phase change in current is produced in each line.

4. That if the array requires equal current drive, driving each element with equal power will always satisfy the requirement.

5. That a current phase angle displacement of 90 degrees between array elements will occur by insertion of a quarter-wavelength line in the feedline of one of the elements.

Every one of these assumptions is wrong, because the premise on which they are based is not true.

Some writers suggest that great liberties may be taken with element feedline lengths. Without considering the effects upon phasing, they would use element feedlines of any length as long as they were equal. Except in very specific circumstances (when all driving impedances are equal), there is no way to justify taking these liberties with most multi-element array configurations.

array impedances and power distribution

It may be illuminating to examine a typical set of dynamic driven impedances for the quarter-wave resonant elements of a 4-square vertical phased array (fed with equal-magnitude currents of the proper phases to produce the main lobe along a diagonal). This will demonstrate the profound effects of mutual coupling.

element 1	$Z_1 = 7.9 - j7.8$
element 2 or 3	$Z_2 = Z_3 = 35.7 - j12.7$
element 4	$Z_4 = 59.2 + j42.6$

The first impedance is the reference, or zero-degree phased element; the next is the impedance of each of the two -90 degree phased middle elements; the last is the -180 degree phased element. That these impedances are quite dissimilar and reactive is obvious. Since drive power is a linear function of the *real* component of these impedances (being fed with currents of equal magnitude), it is clear that power division among these elements is far from equal. Assuming 1-ampere drive to each element, the drive power supplied to each is:

element 1	7.9 watts
element 2	35.7 watts
element 3	35.7 watts
element 4	59.2 watts

which, on a percentage basis, is 5.7 percent, 25.8 percent, 25.8 percent, and 42.7 percent, respectively. Thus a feed network aimed at supplying equal power to this array, such as a Wilkenson power divider, will be at cross purposes with the requirement. (Incidentally, a Wilkenson divider will *not* supply equal power to unequal terminations.) Also, since the 90-degree phased elements are not resistive, simply inserting a quarter-wavelength of delay line in their feeders won't do. Clearly, only a feed system designed for the array elements' driving-point impedances will carry out this unequal power division while producing the proper element phase displacements.

It is possible to devise a feed network which performs these functions while also matching the array to the transmitter feedline. Doing so is not even unduly complex, but calculating the driven impedances does require a knowledge of the self and mutual impedances of the elements. Methods for doing this will be detailed in a future article. The greatest benefit of a good match in multi-element arrays is the warning it provides when loss of continuity to an element occurs because of faulty switching relays or the like.

30 to 40 dB F/B are achievable

My interest in low-band DX began just as described in the beginning of this article. I started with a dipole 30 feet high, then progressed to a vertical, and then to in-line arrays of two and three verticals. With some cut-and-try, the arrays were made to work quite well.

Then came the articles by W1CF on the 4-square

†Special cases are mentioned for completeness. The situations governing them are not ordinarily encountered in phased-array feed network applications. These cases arise when the real and reactive components of a termination have a particular relationship with the characteristic impedance, Z_O, of the line and its electrical length. For example, an eighth-wavelength line will have a current delay of 45 degrees with terminated by an impedance whose arithmetic sum of the real and reactive components equals Z_O. A three-eighths wavelength line, under the same impedance relationships, will exhibit 135 degrees phase delay between input voltage and output current. These are two special cases which I explored; there may be more. I am indebted to W7EL for bringing the possibility of such unusual cases to my attention.

array[4] which inspired me, as they have many others, to duplicate his pathfinding work in building pattern controlled low-band arrays. For me at least, having achieved excellent F/B with simpler arrays (but without bothering to find out precisely why), the F/B results were disappointing. Cut-and-try led nowhere, this array's having too many variables for such blind stabs, and so I had to go back to basics for a more fundamental understanding. Thanks to the advice, encouragement, ideas, and boundless resource of mathematical tools contributed by my friend WB6SXV, as well as many information exchanges with W7EL and W2PV[5] I believe I now know how the 4-square should work.

Achieving theoretical F/B in practice ultimately becomes an exercise in achieving electrical symmetry of the array. This is not easy, but efforts continue to reach that goal. Fortunately, like Yagis, these arrays *want* to work. Less than optimum drive conditions for forward gain find them as tolerant as Yagis, but also as intolerant for high front-to-back ratio. Despite large departures from design drive currents and delay angles, forward gain is not affected much. But seemingly insignificant differences in drive currents or delay angles drastically reduce the maximum F/B capabilities. A 10 percent change in drive current of one element in a 4-square can bring the array from a really excellent 30 to 40 dB F/B down to an average 15 to 20 dB. Another way of looking at this is that excellent F/B ratios hold over a small frequency range, while gain holds over a relatively much larger range, as W2PV showed for the Yagi.[3]

Although the principles for correctly feeding a multiple driven element array have been known since the 1930s,[6,7] their primary application has been by the long-wave a-m broadcast industry, and relatively little has been published in Amateur Radio literature. Perhaps editors may have felt the subject too complex, or that it lacked broad reader interest. Another possible reason is that few modern antenna texts discuss feed methods for such arrays. Typically, many field plots are shown, but means for achieving them are left to the reader.

areas to be addressed

It is the purpose of this series of articles to attempt to fill this gap. Over the next few months I shall try to address the following considerations:

I. Theoretical Array Design
   - Element spacing
   - Drive requirements — magnitude and phase
   - Field plotting — how to calculate

II. Self and Mutual Impedance
   - Measurements and calculations
   - Ground planes
   - Element driven impedances

III. Drive Network Design
   - Four-terminal network matrices
   - Pi and T coax equivalents
   - Directional switching
   - Adjustment and measurement

Topics of this nature cannot be adequately discussed without presenting voltages, currents, and impedances in complex algebraic form, such as R + jX for impedance. Those readers who understand them will have no difficulty in following the presentation; for those who do not, I am assuming that they have a good enough general understanding of the concepts (of resistance and reactance) to be able to understand the implications of the conclusions I present.

In general, I shall try to address myself to general solutions, without restriction to specific designs. Where particular designs are examined, these will be by way of illustration, not for the sake of presenting any one proposal. Rather, it is my hope that readers will find their own solutions to their particular problems within the space they have available. There is nothing writ in stone, for example, which requires the elements of an array to be resonant, to be spaced at 1/4 wavelength, to be phased in multiples of 90 degrees, or to have radials measured to some exact length. Neither do all arrays operate best with equal current magnitude to all elements. A few hours of mathematical experimentation will allow you to run through more designs than you could ever hope to build.

Building vertical phased arrays is not a black art; with accurate measurements of self and mutual impedances and with reasonably good electrical symmetry, theoretical design goals can be closely approximated in practice. Most of the explanation for the large gap between theory and practice which so many builders encounter lies in the many invalid assumptions discussed earlier.

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ham radio

Two crossed dipoles
of different lengths
cover the entire
80-meter band

stagger-tuned dipoles
increase bandwidth

A broadband antenna can be constructed by using a pair of stagger-tuned dipoles, either horizontal or inverted Vs, mounted at right angles to each other and connected in parallel. (See **fig. 1**.) A single 50-ohm coaxial cable is used for the transmission line. The dipoles are of different lengths, with the longer tuned to a frequency near the lower edge of the band and the shorter to a frequency near the upper edge of the band. Because the dipoles are at right angles, no cancellation or nulls occur in the combined radiated field. Near mid-band, the antenna is omni-directional.

The purpose of this article is to derive the basic equations which apply to the standing wave ratio curve for this antenna. These equations are then used to determine the fundamental relationship between the bandwidth and the SWR.

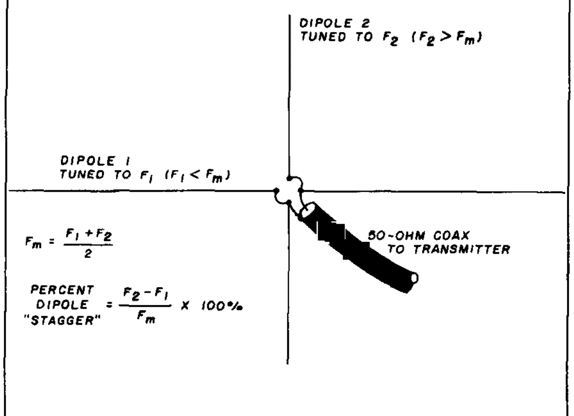

fig. 1. Top view of broadband stagger-tuned, crossed-dipole antenna.

80-meter measurements

The entire 80-meter band is described by a W-shaped SWR curve with a maximum of about 2 (both at the middle and at the band edges). The measured curve of an experimental model is shown in **fig. 2**.[1]

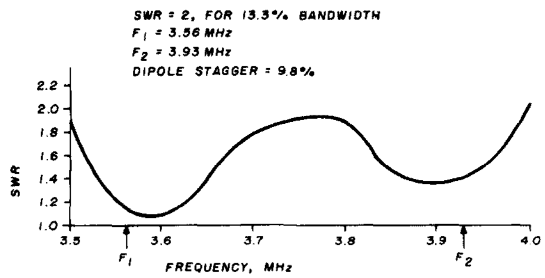

fig. 2. Measured SWR of broadband 80-meter stagger-tuned antenna using inverted Vs.

The 80-meter band, having the greatest percentage bandwidth of all the Amateur bands, is covered by the stagger-tuned antenna without exceeding an SWR of 2. As used here, the term percent bandwidth of a circuit is defined as the bandwidth divided by the mid-band frequency, multiplied by 100.[2] The four Amateur bands considered here, 160, 80, 40, and 10 meters, have bandwidths of 10.5, 13.3, 4.2, and 5.9 percent respectively.

dipole impedance

The stagger-tuned antenna impedance is determined by the impedances of the parallel dipoles. An equivalent schematic for a single center-fed dipole, near its series resonant frequency, is shown in **fig. 3**.

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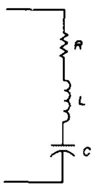


$$Z = R + jX = 66 + 180\left[\frac{F}{F_O} - 1\right] + j1800\left[\frac{F}{F_O} - 1\right]$$

F_O = FREQUENCY OF SERIES RESONANCE

F = VARIABLE FREQUENCY BEING CALCULATED

fig. 3. Dipole equivalent circuit and representative series resonant impedance.

The equation for the impedance of a dipole given in this figure curvature has been derived by fitting to dipole impedance curves.[3]

It is convenient to use normalized impedances, obtained by dividing by 50 ohms, a commonly used coaxial cable characteristic resistance (R_0). At resonance, the normalized radiation resistance of a dipole is 66/50, or 1.32,* numerically equal to its SWR.

For comparison, note that the calculated bandwidth of a horizontal dipole is 7.1 percent at an SWR extreme of 3.0:1. The reactance part of the dipole impedance is about equal to the resistance and the phase angle is approximately 45 degrees. This 7.1-percent bandwidth for a dipole is the basic building block of the stagger-tuned antenna.

impedance of parallel stagger-tuned dipoles

Fig. 4 shows the equivalent circuit for the parallel dipoles, with impedance Z_1 tuned to the lower frequency F_1, and Z_2 tuned to the upper frequency F_2. With F_1 and F_2 fixed, the equation in **fig. 3** first is used for each dipole in turn, to determine the two dipole impedances. From these, and at each frequency, the usual parallel impedance equation of **fig. 4** then gives the stagger-tuned antenna impedance. Finally, the SWR over the entire band is calculated.

For frequencies between the two dipole resonances, the lower F_1 and the higher F_2, an interesting and useful effect exists, which leads to wideband operation. Between F_1 and F_2, the F_1 dipole has a positive reactance while the F_2 dipole exhibits a negative reactance, each being in series with its own radiation (real) resistance. The network acts like a lossy anti-resonant circuit. It is the impedance of this anti-resonant circuit which produces the SWR maximum in the center of the band and limits the attainable bandwidth.

At the center, the two reactances always are equal in magnitude and opposite in sign. The two resistances differ somewhat because of the radiation resistance frequency dependency, that for F_1 being higher than the resistance at its resonance and that for F_2 being an equivalent amount lower. With a further increase in frequency separation (greater than 7.1 percent) the reactances increase faster than the resistances, causing an increasing anti-resonant resistance and SWR.

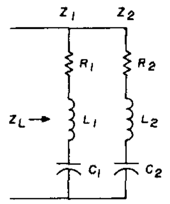
*In general, the series-resonant resistance varies with height above ground and wire size. **Ed.**

calculations

Two W-shaped SWR curves have been prepared, **fig. 5** for an antenna which turned out to be not quite wide enough for the 80-meter band, and **fig. 6** for an antenna not quite wide enough for the 160-meter band. The calculated curves have an appearance remarkably similar to the measured curve of **fig. 2** and confirm that an SWR of less than 2 can be expected for the entire 80-meter band. Using these trial curves, a very good estimate of the needed increase in stagger spacing can be made.

80 meters

The SWR curves of the individual dipoles F_1 and F_2 are drawn in to show that, even when they are far



$$Z_L = \frac{Z_1 Z_2}{Z_1 + Z_2} = R_L + jX_L$$

$$SWR = \frac{1+K}{1-K}$$

$$K = \sqrt{\frac{\left(\frac{R_L}{R_O} - 1\right)^2 + \left(\frac{X_L}{R_O}\right)^2}{\left(\frac{R_L}{R_O} + 1\right)^2 + \left(\frac{X_L}{R_O}\right)^2}}$$

fig. 4. Equivalent circuit of two stagger-tuned crossed dipoles and SWR equation.

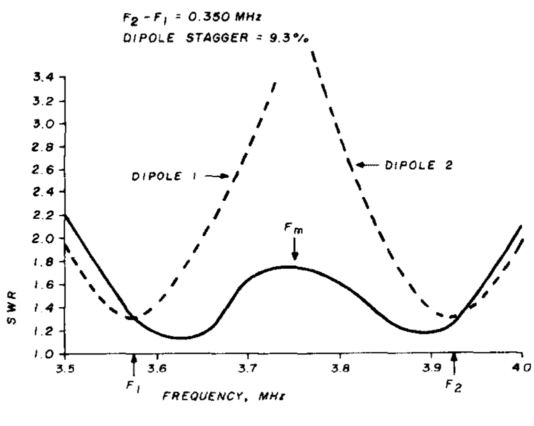


fig. 5. 80-meter stagger-tuned antenna calculated SWR versus frequency.

apart, they still interact to produce an acceptable SWR in the center. Further, the two frequencies where the stagger-tuned SWR curve is lowest, are lower than for individual dipoles.

For the 80-meter band, the stagger should be increased to 10 percent instead of 9.3 percent used in the computations, to fully cover the band. The SWR at the ends and the central maximum is about 2.

160 meters

For the narrower 160-meter band, the stagger should be increased to 9 percent, instead of the 7.9 percent used in the computations. The SWR is about 1.7, less than for the 80-meter band.

10 and 40 meters

The much narrower antenna bandwidths for the 10- and 40-meter bands do not exhibit a center frequency SWR maximum. Instead, a different consideration controls the dipole stagger. The computed SWR curve is shown in **fig. 7** for this condition. As the dipole stagger is reduced to fit these bands, the central maximum disappears and a broad minimum appears. This change occurs at a dipole bandwidth spacing of 7 percent. However, as the minimum is further reduced, the outside edge SWR begins to increase. A choice of 6 percent relative dipole stagger for both the 10- and 40-meter bands appears to be a reasonable compromise.

There is a low overall SWR. This places the dipoles at the edge of the 10-meter band but outside the edges of the 40-meter band! For 10 meters, the maximum SWR is about 1.5, and for 40 meters about 1.3.

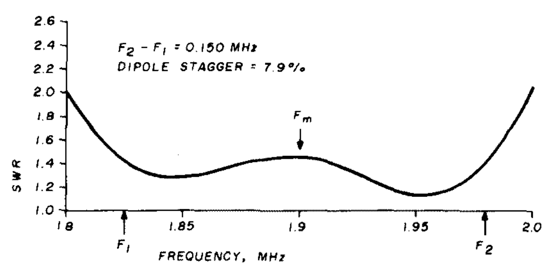


fig. 6. 160-meter stagger-tuned antenna calculated SWR versus frequency.

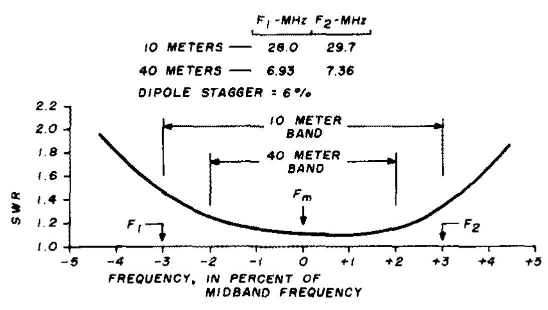


fig. 7. 10- and 40-meter stagger-tuned antenna calculated SWR versus frequency.

summary

Equations which apply to the stagger-tuned crossed dipole antenna have been specified. The 80-meter Amateur band is, relatively, the widest. Measurements and calculations confirm that this entire band can be covered with a SWR of about 2.

A tabulation of the calculated (required) resonant frequencies for the two dipoles, and the calculated maximum SWR for the 160-, 80-, 40-, and 10-meter bands, are given in **table 1**, below:

table 1. Calculated resonant frequencies and maximum SWR.

band meters	F_1 MHz	F_2 MHz	percent bandwidth	calculated maximum, SWR
160	1.81	1.98	9	1.7
80	3.56	3.94	10	2.
40	6.93	7.36	6	1.3
10	28.00	29.7	6	1.5

Note that, for the 40-meter band, the dipole resonant frequencies lie outside the Amateur band. Using only the formula for length is satisfactory, without a direct measurement of the resonant frequency, because the 40-meter band uses only the central 4.2 percent of the antenna's basic 6 percent width.

All the data presented in this article, except the measured curve of **fig. 2**, have been calculated using representative impedances for a dipole.[3] Calculations have insured that the results are comparable throughout, and help determine effects that might not be noticed using only measurements.

A dipole's impedance depends in part on nearby objects and the height above ground. Inverted Vs add even more variables. However, the calculated results show that the stagger-tuned antenna can be adjusted to develop the required wideband characteristic.

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ham radio

20-meter mobile vertical

CB plus replacement auto antenna combine to make a durable performer with good bandwidth

It's unusual to find a homebrew mobile high-frequency antenna, partly because it's rather difficult to construct one which will withstand the 100-plus mph winds antennas encounter from time to time. However, it *is* relatively easy to convert a Radio Shack mobile CB antenna to 20 meters, and the resulting antenna is a surprisingly good performer.

The basis for the antenna is a Radio Shack 4-foot Fiberglass Whip, #21-934 (**fig. 1**). The whip is helically wound near the top and fits a standard 3/8 × 24 threaded mount. When mirror-mounted on my van using a Radio Shack #21-937 mount, the unmodified whip shows an impedance of 25-j1000 ohms at 14.3 MHz. It can be resonated at this frequency by adding, at the top, about 27 inches of straight whip, made from a replacement auto antenna.

Most replacement auto antennas are designed to attach to the broken stub with a set screw. To provide a stub attachment point at the top of the CB whip, carefully scrape away the outer fiber glass material to expose about 1/4-inch of the embedded wire. It appears to be 22-gauge enamel-coated wire. Bare the wire and tin it. Cut the head off a 1/4-inch diameter, 1 1/2-inch brass bolt, and tin the butt (thicker portion) of the threaded end. Solder this end to the wire.

This stub attachment will be secured to the CB whip with glass cloth and epoxy (**fig. 2**). First, however, it's necessary to fasten the bolt to the CB whip to prevent the fine wire from breaking during handling. Lay the CB whip horizontally and block it up so that the auto whip is aligned. Attach the auto whip to the bolt by tightening the set screw, and block it level with the CB whip. Put a dab of 5-minute epoxy on the end of the CB whip, press the two sections together, and visually align them.

After the epoxy has thoroughly hardened, remove the auto whip and sand the top 2 inches or so of the CB whip and bolt with fine sandpaper. This provides a clean surface for the fiber glass reinforcement. Glass cloth/epoxy repair kits are available at most

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hardware stores. Cut three strips of cloth about 12 inches long and 1 inch wide and saturate them with mixed epoxy resin. Starting about 2 inches below the joint, wind each strip in overlapping fashion, like tape, to cover the joint and 1/2 to 3/4 inch of the bolt. Three layers of cloth will provide ample strength. Be sure to follow all instructions provided with the kit, including the safety precautions.

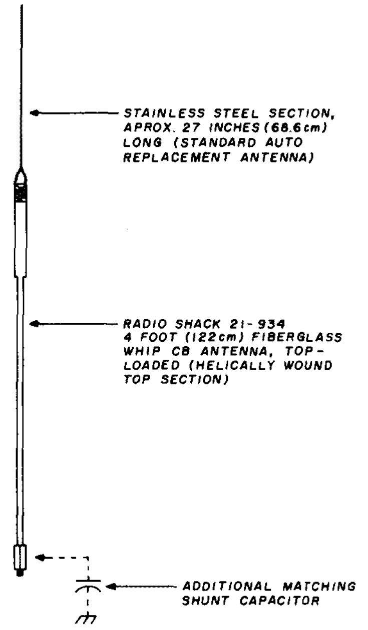

fig. 1. The assembled 20-meter mobile antenna. The capacitor may be attached with spring clips for improved matching (see text).

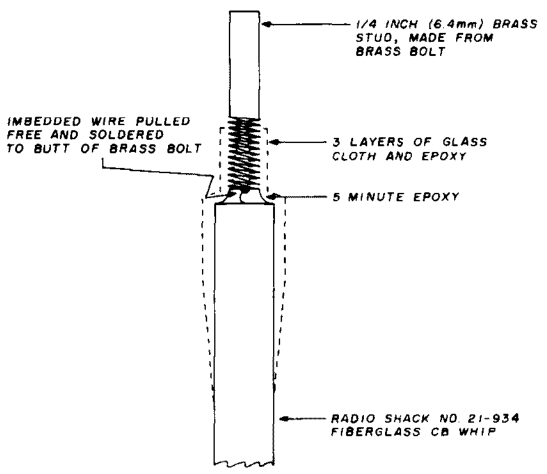

fig. 2. Detail of the attachment point for the replacement auto whip. Use 5-minute epoxy to cement the stud to the whip prior to applying glass cloth.

The joint will be messy at this point, but it can be smoothed out by sanding after the epoxy has cured completely. It probably will be necessary to scrape or sand the exposed portion of the bolt to remove epoxy that has formed on it. Attach the replacement auto whip securely, and you will be ready for tune-up.

matching to the whips

Each mobile installation differs by vehicle, the type of mount, and the position of the antenna on the vehicle. The 27-inch whip length mentioned earlier may not be correct in any installation but mine, but it should be close. Make adjustments as necessary for your situation.

A few hints may make the tune-up process a little less frustrating: 1) use a feedline that is a multiple of one-half wavelength (don't forget the velocity factor) to avoid transformer action in the feedline; 2) park the vehicle well away from objects, such as trees and overhead wires, to prevent detuning; and 3) shorten the whip a *little* at a time — no more than 1/8 inch per cut when you're near resonance. The *ARRL Antenna Book*[1] details other methods for resonating a mobile antenna.

At resonance, the antenna shows a resistive impedance of 30 to 35 ohms. The resulting SWR is adequately low for many purposes, but the bandwidth may be improved by better matching. Trim the whip to be inductive (too long), and then add the appropriate value of shunt capacitance from the base of the antenna to ground, forming an L-network which transforms the impedance to 50 ohms. I found that merely attaching a 200-pF, 500-volt mica capacitor to the base of the antenna and ground with spring clips, then shortening the straight whip until the antenna resonated, provided a feedpoint impedance that was very close to 50 ohms.

When tuned in the above fashion, the antenna showed a 2:1 SWR bandwidth of about 50 kHz for my installation. Consistently good signal reports have come from all areas of the U.S.; I've been running 100 watts PEP (no DX has been attempted). The small-gauge wire used in the CB whip probably has more ohmic loss than desirable, but the performance is not harmed noticeably.

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ham radio

repeater antenna beam tilting

A four-pole collinear reduces a shadowing effect common to mountainous areas

During recent years I've helped design and construct several commercial and Amateur repeaters. Most of these repeaters are located on high mountains where large elevation differences exist between mobile stations and the machine. From such sites, conventional antennas may overshoot the intended coverage area. I wish to introduce a method of electrical beam-tilting which will optimize the use of the antenna radiation pattern.

shadowing and overshoot

Western Montana and the Rocky Mountain region in general have similar topography. In these areas mountains rise from the prairie and valleys to form natural towers for prospective repeaters. Many exceed 10,000 feet (3049 meters) in elevation. From the early years of two-meter repeatering, Montana Amateurs have made use of these sites. In situations like this where very high repeater sites are used, a problem *called* shadowing can exist.

Fig. 1 is an example of a spot where a repeater site is 3000 feet (914 meters) higher in elevation than the desired coverage area. Additionally, the rise in elevation takes place over the relatively short, horizontal distance of five miles (eight kilometers). **Eq. 1** is used to calculate a depression angle of -6.5 degrees from the repeater to the coverage area. (The angle is approximate because curvature of the earth was not included.)

$$\theta = \tan^{-1} \frac{C - R}{D} \qquad \textbf{(1)}$$

where: θ = *The depression angle (degrees)*

R = *The repeater elevation (feet or meters)*

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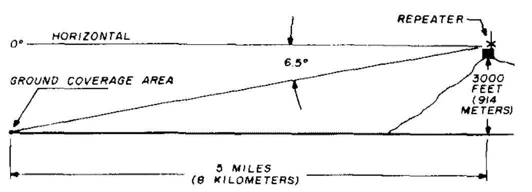

fig. 1. The repeater is much higher in elevation than the intended coverage area. Neglecting the earth's curvature, the antenna radiation pattern must be depressed 6.5 degrees to prevent shadowing of the desired coverage area.

C = *The coverage area elevation (feet or meters)*

D = *The horizontal distance from the repeater to the coverage area (feet or meters)*

Fig. 2 depicts the same site with an antenna radiation pattern added. With the pattern centered on the horizontal or zero-degree line, an antenna radiating a half-power beamwidth of 13 degrees is required for the lower half-power point to fall on the area of desired coverage. Any station closer to the repeater is not in the main pattern and, as a result, is shadowed. Also, the radiated power above the horizontal serves only to heat the ether. If this wasted radiated power could be salvaged to fill in the shadowed areas, a more efficient antenna system would result. I borrowed a solution to both shadowing and efficient energy usage from the field of broadcasting.

the solution

Commercial fm and television transmitters are often located on mountain-top sites. If necessary, beam tilting is used to direct the radiated energy downward from the transmitter site to the intended area of coverage. Beam tilting may be required at lower elevations than one might expect. With consideration given to the curvature of the earth, for example, a transmitter site only 1000 feet (305 meters) above the earth requires a 0.5 degree downtilt for the center of the main antenna radiation pattern to intersect the horizon![1] Obviously, the repeater described earlier could be a serious candidate for beam tilting.

antenna considerations

Through experimentation, the collinear[2] type of antenna seems to be a superior antenna choice for tall, mountain-top applications. Consequently, the discussion is limited to two types of collinear antennas.

The first is the familiar 24 foot (7.3 meters) high, fiberglass encased collinear, which is easy to mount and performs well. Also, some manufacturers will provide an electrical downtilt to your specifications for an additional charge and a shipping delay. For a repeater group working on a shoe-string budget, however, this antenna is not the most economical. Furthermore, if your site is subjected to icing and frequent high winds, the fiberglass collinear may not survive very well. Any small, internal fracture caused by flexing in the wind may cause an rf diode to form and introduce horrid screeches and howls into the repeater. Such slight defects are magnified where the receiver and transmitter of the repeater are closely spaced in frequency. In such cases, the unusable collinear for repeater applications may oftentimes be retired to satisfactory base station service.

I favor a second type of collinear antenna comprising four dipoles fed in phase. This array, illustrated in **fig. 3**, is commonly called the Four Pole antenna. It is derived from linear array theory[3] and can be used for electrical beam tilting. Some commercial Four Poles use folded dipoles with matching baluns as elements, while others use common dipoles with gamma matching or straight feeds. In any case, the antenna feed impedance should be 50 ohms.

Each 72-ohm cable section with a length equal to an odd multiple of a quarter-wave transforms a 50-ohm termination to 100 ohms at the driving end. The resulting 100-ohm impedances are combined in parallel through tee connectors to produce 50-ohm resultants. The 72-ohm coax harness shown in **fig. 3** is used to combine the element impedances to a common 50-ohm feedpoint. Also, since the signal must travel an equal distance from the feedpoint to each element, the elements are fed in phase.

All *cable* length calculations are multiplied by the cable velocity factor to obtain actual lengths. The *an-*

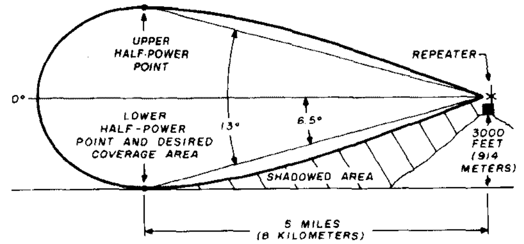

fig. 2. A repeater antenna with a vertical, 13 degree, half-power beamwidth barely meets the required depression angle. Any station closer to the repeater is shadowed while the repeater power in the pattern above 0 degrees is wasted.

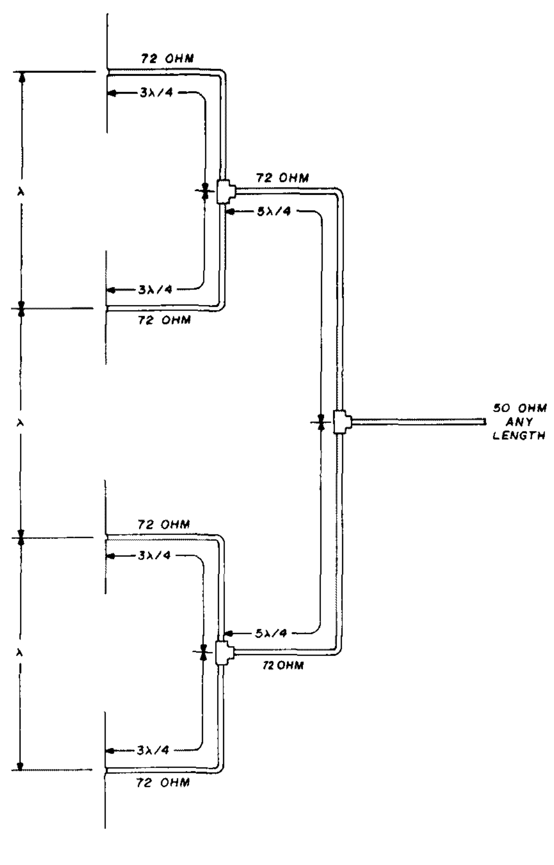

fig. 3. Odd quarter-wavelength coaxial transformers are used to match the dipole impedances to the driving point. The driving phase is the same to all dipoles due to feed system symmetry.

tenna lengths and spacings, however, are close to that of free space.

For strength and mounting convenience, the four dipole elements are usually mounted to a metal mast. Because the mast proximity distorts the element patterns, the four elements should be spaced at 90 degree intervals around the mast to obtain an omnidirectional pattern. If all the elements are mounted on the same side of the mast, the resulting pattern strongly favors the direction the elements are facing. Typical gains for the Four Pole are 6 dB for an omnidirectional pattern or 9 dB for a favored direction. The ability to steer elements and favor specific directions further enhances the utility of the Four Pole as a repeater antenna.

downtilt theory

The vertical radiation pattern of the Four Pole antenna results from pattern multiplication. The Four Pole elements are first considered as a vertical stack of four isotropic radiators (small radiating spheres), each spaced one wavelength above the next. The normalized far field pattern for this linear antenna array is given by **eq. 2**[4] and tabulated in **table 1**.

$$E_a = \frac{Sin\ n\ \ (180^\circ s)cos\,\theta\ +\ \frac{d}{2}}{n\ Sin\ \ (180^\circ s)cos\,\theta\ +\ \frac{d}{2}} \qquad \textbf{(2)}$$

where: E_a = *Field strength of the array (normalized to unity)*

n = *The number of antenna elements*

s = *The antenna spacing from center to center (wavelengths)*

d = *The progressive difference in phase shift between antennas. The top element considered at 0 degrees for reference (degrees)*

θ = *The counterclockwise angle off vertical formed by a line from the array center to the desired field point (degrees)*

(There are a few values that when substituted in **eq. 2** produce an indeterminate form, e.g., zero divided by zero. This problem is overcome by use of the mathematical technique known as L'Hospital's rule* *or* by recalculating **eq. 2** using a slightly greater angle, e.g., $\theta + 1$.)

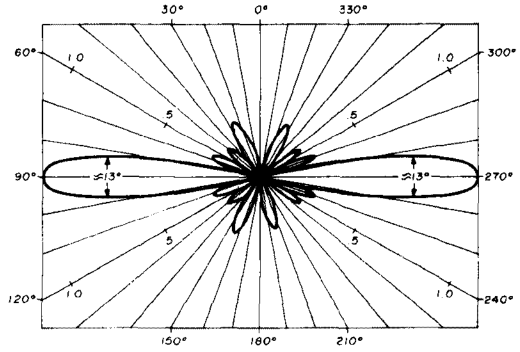

fig. 4. The vertical radiation pattern of a Four Pole antenna produces a half-power beamwidth of approximately 13 degrees. (Half-power points are synonymous with a 0.707 normalized field voltage.) The line of the Four Pole array is centered along the 0 to 180 degree axis with the top dipole in the 0 degree direction.

*Differentiate both the numerator and denominator and substitute 0 for θ. If eq. 2 is still indeterminate, repeat process. Ed.

Once the array pattern is solved the isotropic sources are replaced by vertical dipoles. This is accomplished mathematically by multiplying the linear array pattern by the dipole pattern given in **eq. 3.**[5] (The same method of overcoming indeterminates may be used here as was suggested for **eq. 2.**) The dipole calculations and pattern multiplication results are also shown in **table 1**. The resulting Four Pole pattern is plotted in **fig. 4**.

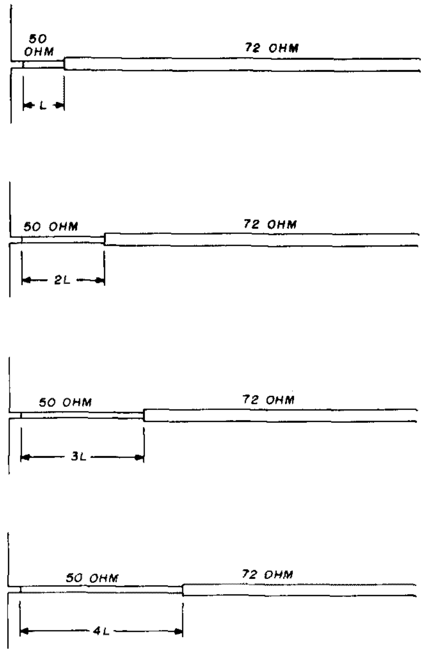

fig. 5. By inserting 50-ohm coaxial delay sections between the dipole feed points and the 72-ohm feed sections, a progressive phase shift is introduced. Selection of the proper phase shift causes the antenna pattern to electrically tilt downward.

$$E_d = \frac{\cos(90^\circ \cos\theta)}{\sin\theta} \quad \textbf{(3)}$$

where: E_d = *Field strength of the dipole (normalized to unity)*

θ = *The counterclockwise angle off vertical formed by a line from the center of the dipole to the desired field point (degrees)*

Since the dipole pattern is a constant, the only hope of creating a downtilt is by modifying some parameter in the linear array pattern. In examining a

table 1. Calculated, normalized field strengths for the linear array (E_a), dipole (E_d), and Four Pole ($E_{Four\ Pole}$) antennas.

Angle θ (Degrees)		$\lvert E_a \rvert$	$\lvert E_d \rvert$	$\lvert E_{Four\ Pole} \rvert$
1,	181	1.000	.014	.014
11,	191	.992	.151	.150
21,	201	.894	.291	.260
31,	211	.562	.432	.243
41,	221	.021	.573	.012
51,	231	.272	.708	.193
61,	241	.048	.828	.040
71,	251	.238	.922	.291
81,	261	.489	.982	.480
83.5,	263.5	.710	.991	.704
91,	271	.993	1.000	.993
96.5,	276.5	.710	.991	.704
101,	281	.300	.973	.292
111,	291	.271	.906	.246
121,	301	.047	.805	.038
131,	311	.262	.682	.179
141,	321	.130	.545	.071
151,	331	.651	.404	.263
161,	341	.928	.263	.244
171,	351	.996	.124	.120

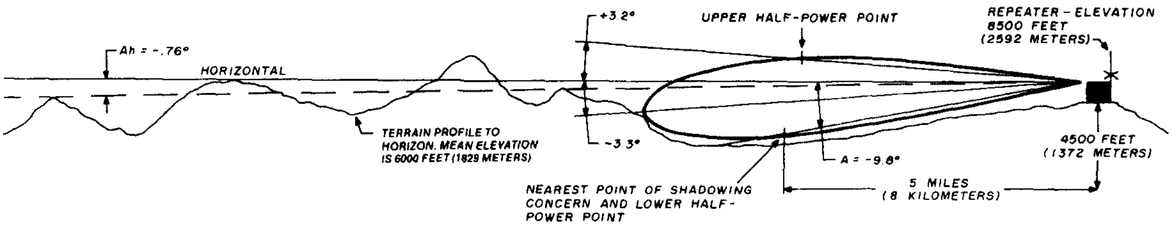

fig. 6. By selecting the proper beam tilt through the information presented in the text, shadowing may be eliminated through maximum utilization of the repeater antenna pattern. The curvature of the earth and the angle from the repeater to the horizon must also be considered.

table of linear array patterns,[6] I discovered that a progressive phase delay of the lower Four Pole elements would, in theory, tilt the beam downward. In practice, the phase delay is accomplished by placing appropriate lengths of 50-ohm coax cable between the antenna element feed points and the 72-ohm phasing harness, as illustrated in **fig. 5**. Although it is not necessary to add any length to the top element, doing so overcomes any phase errors caused by the addition of connectors used in the lower element phasing sections.

determining downtilt

The first step in formulating a downtilt is to determine two depression angles. The first is the angle of the horizon and the second is the deepest angle where repeater shadowing is *not* to be allowed. **Eqs. 4** and **5** may be used for these angles respectively.[7]

$$A_h = \frac{0.0108P}{D} \tag{4}$$

where: A_h = *The depression angle to horizon (degrees)*

P = *The elevation difference of the repeater site over the average terrain elevation (feet — multiply meters by 3.28)*

$$A = \frac{0.0109\,H}{D} \tag{5}$$

where: A = *The depression angle (degrees)*

H = *The elevation difference of the repeater site over the nearest point of shadowing concern (feet — multiply meters by 3.28)*

D = *The horizontal distance from the repeater to the nearest point of shadowing concern (miles — multiply kilometers by .62)*

Once these angles are known, the half-power beamwidth of the Four Pole may be fitted to these angles to provide optimum coverage.

For example, assume a repeater is located on an 8500 foot (2591 meter) peak overlooking a valley with an elevation of 4000 feet (1220 meters) and the difference in elevation occurs over a distance of five miles (eight kilometers). Because the area is fairly mountainous, the average terrain height to the horizon could be estimated at 6000 feet (1829 meters). From **eq. 4** and **5**, the two depression angles are $A_h = 0.76$ *degrees* and $A = 9.8$ *degrees*.

Fig. 4 and **table 1** show the half-power beamwidth of the Four Pole to be very close to 13 degrees or 6.5

```
20 CLS:PRINT:PRINT
30 PRINT"                    * * * * ANTENNA POLAR PLOTTING PROGRAM * * * *"
40 PRINT:PRINT"                                 BY LEE BARRETT K7NM 1981"
50 PRINT:PRINT"                       * * * * PRESS ANY KEY TO CONTINUE. * * * *"
60 Y=RND(191)
70 FOR X=15360+63 TO 15360 STEP-1:POKE X,Y:NEXT
80 FOR X=15360 TO 16383 STEP-64:POKE X,Y:NEXT
90 FOR X=15361 TO 16383 STEP-64:POKE X,Y:NEXT
100 FOR X=15362 TO 16383 STEP-64:POKE X,Y:NEXT
110 FOR X=15360+960 TO 16383:POKE X,Y:NEXT
120 FOR X=16383 TO 15360 STEP-64:POKE X,Y:NEXT
130 FOR X=16382 TO 15360 STEP-64:POKE X,Y:NEXT
140 FOR X=16381 TO 15360 STEP-64:POKE X,Y:NEXT
150 R$=INKEY$:IFR$=""THEN 150  ELSE 170
160 GOTO 60
170 DIM E(360),M(70,50),Z(360),Y(360):G$="##.#####"
180 CLS
190 PRINT"TYPE NUMBERS GREATER THAN 10 FOR A YES ANSWER AND NUMBERS LESS THAN 1
0 FOR A NO ANSWER TO QUESTIONS."
200 PRINT:PRINT
210 INPUT"INPUT THE NUMBER OF ANTENNAS.";N
220 PRINT
230 INPUT"INPUT THE SPACING IN WAVELENGTH.";D
240 PRINT
250 INPUT"INPUT THE PROGRESSIVE PHASE SHIFT IN DEGREES.";P
260 PRINT
270 INPUT"INPUT DEGREE STEP DESIRED FOR CALCULATIONS.";K
280 CLS
290 C=3.14159
300 R=((2*C)/360)
310 FOR A=1 TO 360 STEP K
320 S=(((2*C*D)*COS(A*R))+(P*R))
330 B=(N*SIN(S/2))
340 IF B=0.0 THEN 370
350 T=SIN((N*S)/2)
360 GOTO 390
370 B=(N*COS(S/2))*((C*D)*SIN(A*R))
380 T=(COS((S*N)/2))*((N*C*D)*SIN(A*R))
390 E(A)=T/B
400 PRINTTAB(8)"ANGLE=";A;TAB(23)"E = ";USING G$;ABS(E(A));:PRINT
410 NEXT A
420 INPUT"PLOT ARRAY?";X
430 IF X > 10 THEN 680
440 INPUT"NEW PATTERN?";X
450 IF X > 10 THEN 180
460 FOR A = 1 TO 360 STEP K
470 B=SIN(A*R)
480 IF B<>0.0 THEN 520
490 B=COS(A*R)
500 T=(-1*SIN((C/2)*COS(A*R)))*((C/2)*(-1*SIN(A*R)))
510 GO TO 530
520 T=COS((C/2)*COS(A*R))
530 Z(A)=T/B
540 PRINTTAB(8)"ANGLE=";A;TAB(23)"E=";USINGG$;ABS(Z(A));:PRINT
550 NEXT A
560 INPUT"PLOT DIPOLE?";X
570 IF X > 10 THEN 720
580 PRINT"FOUR POLE PATTERN:"
590 FOR A=1 TO 360 STEP K
600 E(A) = E(A)*Z(A)
610 PRINTTAB(8)"ANGLE=";A;TAB(23)"E=";USINGG$;E(A);:PRINT
620 NEXT A
630 INPUT"PLOT 4 POLE PATTERN?";X
640 IF X > 10 THEN 680
650 INPUT"QUIT?";X
660 IF X > 10 THEN 1080
670 GOTO 180
680 FOR A=1 TO 360 STEP K
690 Y(A)=ABS(E(A))
700 NEXT A
710 GOTO 750
720 FOR A = 1 TO 360 STEP K
730 Y(A)=ABS(Z(A))
740 NEXT A
750 FOR I=1 TO 64
760 FOR L=1 TO 50
770 M(I,L)=0.0
780 NEXT L
790 NEXT I
800 FOR A=1 TO 360 STEP K
810 U=(Y(A)*32)*SIN(A*R)
820 O=(Y(A)*23)*COS(A*R)
830 U=(31-INT(U))
840 O=(23-INT(O))
850 M(U,O)=1
860 NEXT A
870 CLS
880 PRINT TAB(50)"REFERENCE ANT"
890 PRINT TAB(50)"AT BOT. OF "
900 PRINT TAB(50)"ARRAY. ZERO"
910 PRINT TAB(50)"AT THE TOP"
920 PRINT TAB(50)"WITH CCW"
930 PRINT TAB(50)"INCREASING"
940 PRINT TAB(50)"ANGLES."
950 FOR I=1 TO 63
960 FOR L=1 TO 47
970 IF M(I,L)<>1 THEN 990
980 SET (I,L)
990 NEXT L
1000 NEXT I
1010 PRINT@930,"CALCULATE NEW ARRAY?"
1020 INPUT X
1030 IF X >10 THEN 180
1040 INPUT"CALCULATE DIPOLE?";X
1050 IF X  >10 THEN 460
1060 INPUT"CALCULATE COMPOSITE?";X
1070 IF X  >10 THEN 580
1080 END
```

fig. 7. The program presented is compatible with the TRS-80® microcomputer and may be used to speed and simplify the downtilt design calculations.

degrees from the beam center to either side. If necessary, the half-power beamwidth may be widened by placing the elements closer together (this can be done by substituting proper value of s less than one in **eq. 2**).

At this point, a decision must be made as to where the energy is to be distributed. I decided to use one wavelength spacing and to place the lower, half-power point at a depression angle of 9.8 degrees. As illustrated in **fig. 6**, the beam must be tilted downward 3.3 degrees. The upper half-power point then occurs at an elevation angle of 3.2 degrees which allows some of the signal to bend over the horizon to the DX stations.

Although the calculations may be done by hand using **eq. 2** and **3**, I used a computer program for the TRS-80 which is listed in **fig. 7**. As illustrated in **fig. 8**, the program results indicate that a three-degree depression angle can be achieved with a fifteen-degree progressive phase delay to the lower Four Pole elements. The fifteen-degree phase delay coax length is calculated using **eq. 6**.

$$L = \frac{C}{f} \times \frac{P}{360} \times 100V \qquad (6)$$

where: L = *The phase delay coax length (centimeters — divide by 2.54 for the length in inches)*

C = *Velocity of a wave in free space (300,000,000 meters/second)*

f = *The operating frequency (Hertz)*

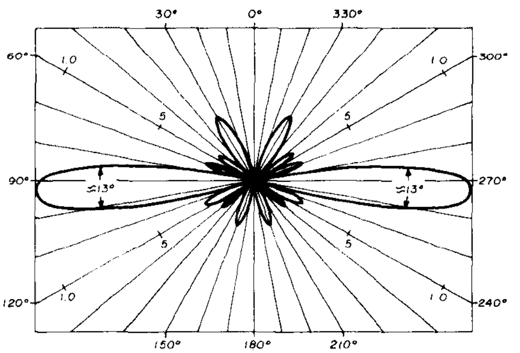

fig. 8. A 15-degree progressive phase delay (with the top dipole having the shortest delay line) produces the indicated beam tilt. The main radiation is depressed nearly 3 degrees from vertical while the half-power beamwidth remains near 13 degrees. The array is centered on the 0 to 180 degree axis with the top dipole in the 0 degree direction.

P = *The phase delay required (degrees)*

V = *The velocity factor of the coax to be used*

The phase delay length for 146.88 MHz will be 2.25 inches (5.72 cm), assuming the velocity factor of the coax to be 0.677. Referring to **fig. 5**, the phase delay coax lengths from top to bottom will be 2.25 inches (5.72 cm), 4.5 inches (11.43 cm), 6.75 inches (17.15 cm), and 9 inches (22.86 cm), respectively.

mechanical considerations

If the Four Pole is to remain free-standing in a high wind and ice environment, the antenna should be guyed at the top. A nonconducting guy cable such as Phillystran® may be used with standard-size cable clamps.[8] The bottom three to four feet (.914 to 1.22 meters) should be steel guy cable to prevent rodent damage.

Having worked on some pretty tough sites, I put quite a lot of thought into a Four Pole that would survive. **Fig. 9** details such an antenna, which I intend to test in the near future. A nonconductive support structure such as a wooden pole is ideal. However, with the antenna spaced a wavelength from the support structure, even a metal tower should not greatly degrade the pattern.

The antenna is constructed of alternate sections of the insulating guy material previously mentioned and no. 10 solid copper. Sections of PVC pipe are used to protect the horizontal runs of the phasing harness coax from ice damage. Since the tensile strength of the cable antenna is large, the antenna is used to support one end of the PVC sections. The opposite PVC section ends are clamped to the support structure.

A turnbuckle is used to tighten the antenna. The cross-sectional area of the antenna is small and presents a very low wind resistance. Any vibration in the cable antenna tends to clear itself of ice. Finally, the antenna is omnidirectional because the elements are truly collinear.

conclusion

Antennas and mousetraps seem to fit the same category — someone is always after a better one. At present, one downtilt system has been tested. From this initial experience, the downtilt seems to reduce the amount of mobile chopping usually experienced in the canyons and gullies. Only one comparative test has been made, and in that test the downtilt was generally better than the standard Four Pole in both transmitting and receiving.

This is an early stage in my experimentation with downtilt antennas and I would appreciate receiving

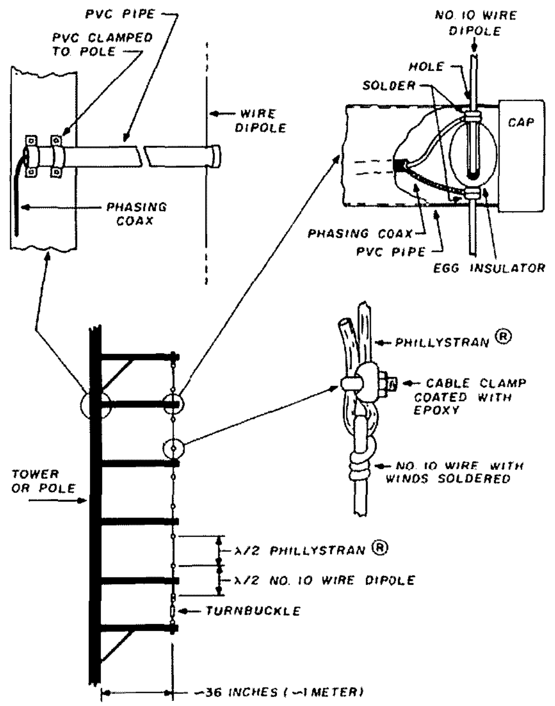

fig. 9. This suggested Four Pole array (constructed of cable with high tensile strength) may solve the tough environmental problems encountered on mountain tops. PVC pipe is used to protect horizontal feed cable runs from ice build-up.

any test results others might gather using these antennas.

acknowledgments

Thanks go to Dennis Nord, WB7UOI, for his aid and assistance with computing and testing of antennas.

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3. John D. Kraus, *Antennas*, McGraw-Hill, 1950, pages 76-83.

4. Ibid, page 78.

5. Ibid, page 142.

6. Carl E. Smith, *Theory and Design of Directional Antennas*, Smith Electronics, 1969, Appendices B-1 & B-9.

7. *AM/FM Radio Station Application Data and Reference Guide*, Broadcast Systems - RCA, 1978, page 44.

8. Phillystran®, a product of Philadelphia Resins Corporation, P.O. Box 454, 20 Commerce Drive, Montgomeryville, PA 18936.

ham radio

short vertical antennas for the low bands: part 1

Relative performance of 5 different shortened verticals is compared to full quarter-wave radiator

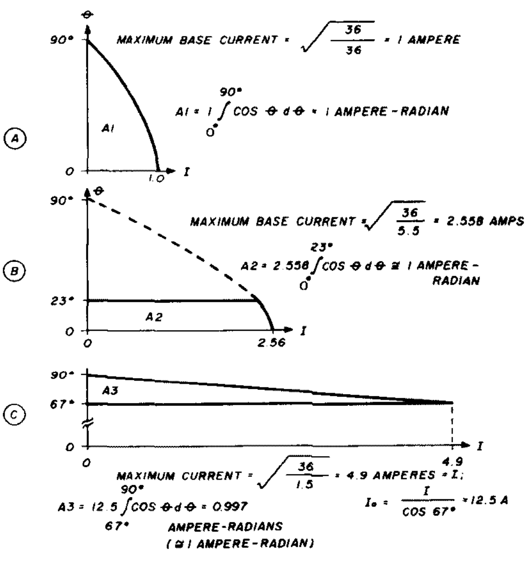

fig. 1. Current distributions for three referenced antennas, each over perfect ground, and 36 watts input to antenna terminals. All to scale, so areas are directly comparable. *(A)* Current distribution of λ/4 vertical (reference antenna) against a perfect ground, *(B)* Current distribution of λ/16 (23 degree) top-loaded vertical against perfect ground, *(C)* Current distribution of λ/16 (23 degree) base-loaded vertical against perfect ground.

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The increasing popularity of the 160-meter band and recent FCC regulatory actions opening the lower 100 kHz to normal Amateur operations have attracted Radio Amateurs to the top band. Many are discovering that wire antennas normally used on the higher frequencies require difficult to achieve heights and lengths for effective operation, especially 160 meters.

The decision to investigate verticals rather than doublets or other horizontal antennas resulted from space limitations and performance requirements. (A maximum height of 35 feet, one of the constraints, equates to 1/8 wavelength on 75 meters and 1/16 wavelength on 160 meters. Most horizontal antennas at this height above *ground* provide only high-angle radiation.) A two-band trapped vertical is described that uses the same radiating element for both bands

and isolates the top-loading capacity hats with a trap. A short vertical can be nearly as efficient as a full-size quarter-wave vertical if it is top-loaded, and has an extensive ground system.

design considerations

A quarter-wave vertical has a radiation resistance of approximately thirty-six ohms.[1] In quarter-wave (or shorter) systems, over non-ideal ground, a total resistance (R_T) would be:

$$R_T = R_r + R_\Omega + R_g$$

where R_r = radiation resistance
R_Ω = circuit resistance
R_g = ground resistance

Fig. 1 illustrates the calculated current distribution for three verticals. **Fig. 1(A)** is a plot of the current in the perfect quarter-wave, **fig. 1(B)** for a 23-degree high, top-loaded vertical, and **fig. 1(C)** for a 23-degree high, base-loaded system. **Figs. 2** and **3** show the values for helical, center-loaded, and 50/50 top- and base-loaded verticals, all 23 degrees in electrical height. The calculations show that short verticals can be nearly as efficient as full-size antennas. (The 23 degree electrical length is related to my height restriction.)

Short antennas have current distributions that can be approximated by triangular or trapezoidal shapes. The set of curves illustrated in **fig. 4**, extrapolated from a standard reference volume on antenna design[2] are used to determine the radiation resistance of short verticals for defined current distributions.

The curves worked very well for the 160-meter version of my antenna. I departed from the specific domain of the curves in the evaluation of the radiation resistance of the 75-meter system. The 19-ohm resistance for a top-loaded 48.9-degree-high vertical (determined from **fig. 4**) is very close to the measured value and to the value derived by original methods. **Figs. 5** and **6** resulted from my not knowing how far (or whether) to extrapolate the curves in **fig. 4**. **Fig. 5** has been modified to fit two well-measured resistances, but it is within three to five percent on the curve as derived. As modified, it is probably within one percent anywhere for θ between 3 and 90 degrees. **Fig. 6** presents the radiation resistances of base-loaded verticals ranging from 6 degrees to 90 degrees in height. Other combinations of base-loading and top-loading result in radiation resistances somewhere between these curves.

Free-space wavelengths were used to calculate antenna heights. No attention was given to the element length-to-diameter ratio, or to end-effects. For most systems the length-to-diameter ratio is high, and the differences between, say 20 degrees and 21 degrees in terms of radiation resistance is negligible.

Once the calculations were made for the radiation resistances, the feedpoint resistances were defined, and the final evaluation proceeded.

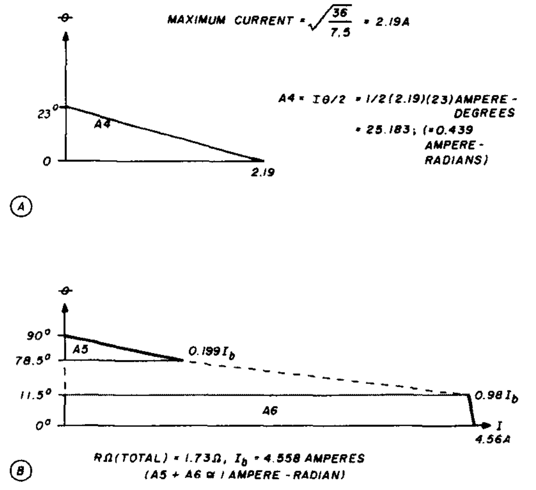

fig. 2. Current distributions for λ/16 (23 degree) helical and λ/16 center-loaded vertical antennas over perfect ground. No coil loss is assumed in center-loaded case, but 6-ohm helical ohmic resistance was included in *fig. 2A*. Figure is drawn to same scale as *fig. 1*. *(A)* Current distribution of λ/16 (23 degree) helical vertical over perfect ground, but with 6 ohm helical resistance. *(B)* Current distribution of λ/16 (23 degree) center-loaded vertical over perfect ground, no coil loss.

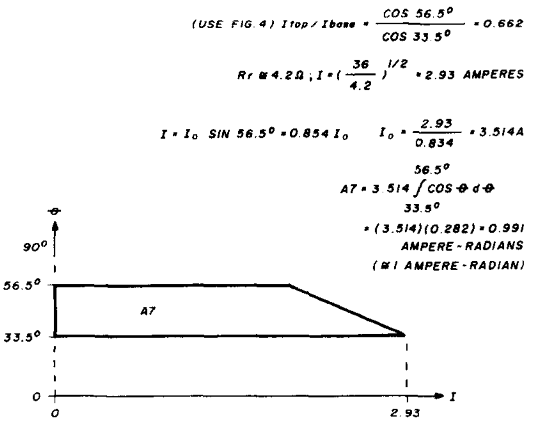

fig. 3. Current distribution on λ/16 (23 degree) vertical antenna with equal top- and base-loading, over perfect ground, no coil loss. (Same scale as *figs. 1* and *2*.)

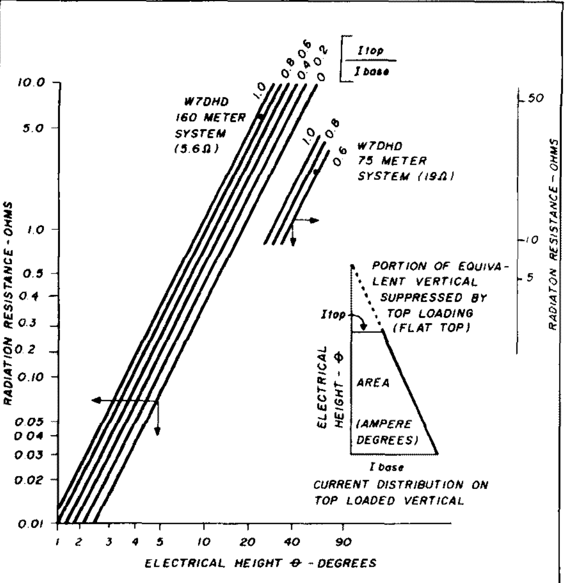

fig. 4. Radiation resistance versus angular aperture (electrical height), θ, for top-loaded vertical antennas.[2]

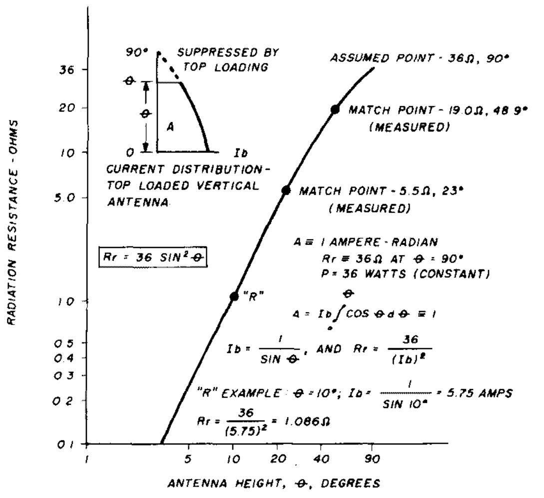

fig. 5. Radiation resistance of top-loaded vertical antennas from 3 to 90 degrees. Theoretically derived and modified to fit measured resistances (Δr less than 1 ohm at $\theta = 50$ degrees).

In all calculations a lossless quarter-wave vertical was used as reference. Field strength is directly proportional to the product of the length of radiating element, and the current in that element in ampere-degrees or ampere-radians. The areas under the profiles of currents in **figs. 1** through **3** are equal to one ampere-radian for 36-watts of input power. The one exception, the helical antenna, was calculated at six ohms rf-resistance in the helicoid, and the integration was done graphically, since current varies linearly along its length.

evaluation

In order to compare the vertical antennas, a ground system consisting of 40 1/8-wave radials was used.

A quarter-wave vertical working against this ground system (12 ohms at 1.8 MHz) exhibits a 75-percent efficiency.[3] This ground system is now used with the shortened verticals.

Since the calculated radiation resistance for a $\lambda/16$ base-loaded vertical is 1.5 ohms (see **fig. 6** with $\theta = 23$ *degrees*), the efficiency is

$$\eta = \frac{1.5}{1.5 + 12 + 2} = 9.7 \text{ percent}$$

where the 2 in the denominator is the rf resistance of the wire in the base-loading coil. Consequently a base-loaded antenna over the same ground system is one-tenth as efficient as a lossless quarter-wave antenna.

Since efficiencies are indicative of radiated field strengths, signal levels, referred to the quarter-wave standard, would be:

$$20 \log_{10} (\text{relative efficiency}) = dB$$

In the case of the base-loaded vertical, this becomes:

$$20 \log_{10} (0.097) = -20.26\ dB$$

Table 1 lists the expected performance of seven vertical antennas:

All the calculations are the same, with the exception of the helical vertical. It was evaluated by making some assumptions: it requires $\lambda/2$ of wire to achieve $\lambda/4$ resonance; wire size is No. 12, 250 feet, $R_\Omega = 6$ ohms; overall height is 35 feet, or 23 degrees; very small (<1 degree) top-hat (the pie tin); the current decreases linearly over the helix.

The current distribution is triangular with an area equal to 1/2 $I\theta$ ampere-degrees. It ranks seventh out of seven verticals, and was not further considered. It is a poor choice, especially when the amount of material and the difficulty of construction are considered.

actual design

Two-band operation would be achieved with the same radiator if a method of switching top hats could be engineered. This was accomplished by use of two separate top hats and a parallel-resonant trap.

table 1. **Relative ranking of several vertical systems by field strength, constant 23 degrees aperture and constant power input.**

antenna system	description	conditions	relative field strength, dB
A	full-sized λ/4 vertical	zero losses	0
B	full-sized λ/4 vertical	12 ohm ground	−2.5
C	λ/16 top-loaded	12 ohm ground	−10.0
D	λ/16 top and base loaded	12 ohm ground, 1 ohm coil	−12.4
E	λ/16 center-loaded	12 ohm ground, 2 ohm coil	−19.25
F	λ/16 base-loaded	12 ohm ground, 2 ohm coil	−20.26
G	λ/16 helical	12 ohm ground, 6 ohm coil	−20.28

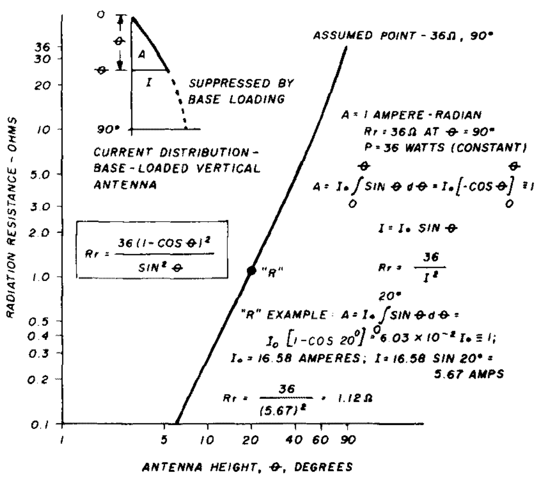

fig. 6. Radiation resistance of base-loaded vertical antennas from 6 to 90 degrees high, (theoretically derived).

The 75-meter top-hat seemed achievable while top-loading on 160 meters (accounting for 67 degrees or 105 feet of missing vertical) seemed more formidable. One source in 1915[4] describes short vertical antennas that use umbrella-loading for top hats.

trap affects performance

Antenna performance depends on the behavior of the trap, tapped onto the 75-meter section at the 49 degree point. The voltage is estimated at 1200 volts, peak, at a one-kilowatt power level. Since the large umbrella is connected to the other side of the trap, that end is assumed to be held constant at or near zero potential. The entire voltage appears across the trap.

The T-200-2 (red core) powdered-iron toroids were wound with No. 12 solid copper wire and resonated with 400 pF at 3.8 MHz. The fundamental wave shape was observed at the kilowatt level for signs of distortion and for ticks in the reflected power on the Bird wattmeter. This was done to determine whether the trap core saturates. No calculation was performed during design — an oversight.

The trap is subjected continuously to the same abuse as is a tank circuit of a kilowatt linear which is unloaded, dipped to resonance, and driven by an exciter. Any trap must be designed to withstand that treatment. Consequently, any trap in any system should be built from the same size and quality components used in the amplifier that drives them — preferably better quality.

power dissipated in the trap

With a trap-resonating capacitance of 400 pF, and a trap-inductance of 4.5 μH, both exhibit 108 ohms at 3.8 MHz, while the ten feet of No. 12 wire has an rf resistance of 0.25 ohms. This calculates to 31 watts of power, dissipated by the trap. This would prove very significant if the antenna were subjected to five or ten minutes of RTTY or a-m operation.

These considerations must be balanced by other factors. If the trap Q is increased, the loss is reduced; but so is the system bandpass. These are engineering trade-offs. The trap in this system effectively limits the 75-meter bandpass (between 2:1 VSWR points) to 86 kHz. Other methods are used to circumvent that limitation.

Another characteristic of short antennas is their very low feedpoint impedance — so low that it is sometimes hard to measure. In highly efficient systems the inclusion of even one ohm of non-radiating resistance will make a significant change in the feedpoint resistance. The equivalent series-input resistance (R_Ω) of the trap resistance, calculated above,

may be estimated very closely if the as-built base current is known:

Given $P_D = 30.9$ *watts (dissipation in trap)*
and $I_B = 7.14$ *amperes*

then $R_{eq} = \dfrac{30.9}{(7.14)^2} = \dfrac{30.9}{51} = 0.61$ *ohms*

So it is known already that the trap with its 0.25-ohm coil resistance will be reflected at the antenna base as 0.61 ohm in series with the other instrinsic resistances.

The calculated radiation resistance for the 75-meter system is 19 ohms. The measured feedpoint resistance is 19.6 ohms. It is highly probable that the 0.6-ohm discrepancy can be explained by the rf resistance of the trap, calculated in the preceding paragraphs.

The construction, measurements, and performance characteristics of verticals in general, and of a two-band trapped vertical antenna in particular, will be described in Part 2, the conclusion of this article.

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3. Jerry Sevick, "Short Ground-Radial Systems for Short Verticals," *QST*, April, 1978.
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ham radio

handi-antennas

Three antenna improvements for 2-meter hand-helds — including a backpack beam!

No Amateur Radio station is better than its antenna. Most antenna articles seem to be for DXers or big-gun contesters, but this article discusses antennas for use with HT's that you can literally hold in your hand. The first is a mechanical improvement in attaching the rubber duckie to an HT; the second is an antenna with significant gain over the rubber duckie; the third is a backpack beam..

a mechanical improvement

One day I had a QSO with WD6FMG who had just finished replacing the output connector on his HT. He remarked that he had been in the habit of removing the rubber duckie every time he put the HT in his briefcase or connected the mobile whip. After removing and replacing the antenna many times, the connector had worn down and would not make a reliable contact. It was a difficult job to disassemble the transceiver and replace the output connector.

This started me thinking. Wanting to avoid the same problem, I looked for a sacrificial connector, an adapter that could take the wear and then be replaced easily. My first approach was to use a straight male-to-male adapter and a straight female-to-female adapter. This served the purpose, but seemed cumbersome.

Then I discovered that a BNC 90-degree elbow had one male and one female end. This could be used as a sacrificial connector, but made it necessary to hold the HT on its side, which is awkward.

It's best to use two right-angle adapters. This arrangement has several benefits in addition to the sacrificial adapter. It is not necessary to remove the rubber duckie to put the HT in a briefcase, because it folds down compactly. For mobile or other use with an external antenna, the two adapters act as a swivel. This allows the antenna connector to bend and rotate when the HT is picked up or set down.

a performance improvement

Despite the convenience of the rubber duckie, there are many times when an antenna with more punch is needed. This is particularly true when operating simplex, or in populous areas where repeater sensitivity must be restricted so that high-power stations do not bring up several machines at the same time.

Rubber duckie antennas are not particularly efficient; a quarter-wave whip can achieve 3 to 6 dB more radiated signal than the rubber duckie. Considering the threshold effect of fm, 3 dB can make all the difference between good copy and no copy at all.

Several more dB can be achieved over the quarter-wave whip by paying attention to the image or ground side of the antenna. There have been several articles describing the importance of a proper ground structure in achieving a low angle of radiation. After all, low angle radiation is the name of the game to increase your coverage on 20 meters or 2 meters.

The easiest way to provide the ground side of the antenna is to use another quarter-wave whip positioned down from the antenna feed point. The result is really a center-fed vertical dipole. I made this type of antenna as an experiment, and was very pleased with the improvement in signal strength for such a simple design. I have since used the dipole antenna

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The station, ready to be assembled. The longer antenna parts are in a plastic bag. The smaller pieces and a length of coaxial cable are in the smaller pouch. (Photo courtesy N6ST.)

to make contacts that would have been impossible with only a quarter-wave antenna. The half-wave dipole seems to be competitive with a five-eighths whip without the problem of damaging the ceiling. Of course, you can reduce the half-wave to a quarter-wave when signals are strong enough.

The half-wave antenna is basically two quarter-wave whips and a BNC tee adapter. The only trick is that the two whips must be fed out of phase: one whip connects in the normal fashion to the center conductor; the other whip is electrically connected to the outside of the connector. This is just like an 80-meter dipole connected to coax cable, where one side of the antenna is connected to the cable shield.

One way to make the required connection is to modify the tee fitting so the center conductor on one end connects to the outside of the connector instead of the other center conductors. This procedure should take only a few minutes. Carefully drill a small hole, about 1/8-inch (0.32-cm) diameter, slightly closer to one end than the other. Drill through the center conductor close to where it joins the center conductor from the side arm. If you are uncertain precisely where to drill, it may be preferable to enlarge the hole in the connector shell. Then use a pointed knife blade to cut away the plastic and expose the three center conductors where they join. Next drill through the center connector going to one end. Be sure to leave the remaining two center conductors joined.

The next step is to remove the piece of center conductor from the connector. If you are lucky the piece may drop out, but the drill probably will have created enough of a burr to hold the piece in place. Insert a short length of 18-gauge wire into the center contact. This will allow a pair of tweezers to grasp the contact without damaging the contact fingers. Gently pull the cut center conductor out of the end of the connector.

The piece of 18-gauge wire can serve as a handle while you perform the next steps on the piece you just removed. Trim this piece a little bit shorter so there will be a gap when it is re-installed. Solder a short piece of flexible wire to the end of the cut piece.

You are now ready to reassemble the connector. Thread the wire and attached piece of center conductor back into the connector. Solder the end of the wire to the outside of the connector. Use an ohmmeter to verify that the rewired center conductor is connected to the outside of the connector and not to the other center conductors. Fill the hole with epoxy to provide mechanical support for the rewired center conductor. When the glue hardens you are ready to try it out.

A completely different approach is to use a standard tee fitting, one standard whip and one modified whip. The connector of the modified whip has its center pin and insulating spacer removed. The insulator is replaced by a solid metal piece so the whip connects directly to the shell of its connector.

still more gain

My next design objective was to design and build a 2-meter antenna with 10-dB gain which could be folded or disassembled into a size not more than 16 inches (40 cm) long: small enough to fit into a backpack. But what kind?

table 1. Gain of driven arrays.

number of elements	possible gain
1	0 dB (reference)
2	3 dB
4	6 dB
8	9 dB
16	12 dB

How much gain can you achieve with a driven array of dipoles? Adding a second dipole to the reference antenna can add up to 3 dB to the gain figure. Another 3 dB is achieved by adding two more dipoles to the array, for a total of four elements. **Table 1** summarizes the number of elements required for a driven array of given gain, and shows why a driven array of dipoles is not attractive for use as a portable antenna.

This brings us to another category of antennas, parasitic arrays. **Table 2** shows the gain you can expect from a properly designed Yagi or quad using a reasonable number of elements. From this, we can expect to get 10-dB gain over a dipole from either a four-element Yagi or a three-element quad. This is much more promising than a ten-element driven array. In fairness, it should be pointed out that driven arrays can generally be made to work over a broader range of frequencies than parasitic antennas. Also, a quad or Yagi will require rotation toward the station, while a colinear antenna, having an omni-directional pattern, does not.

Quads are great antennas. I used a full-size quad on 20 meters for many years. Quads can also be mechanical marvels (or monsters depending on your point of view). The challenge is to build a bigger antenna that packs smaller! Quad antennas usually have mechanical spreaders which support the elements. In contrast, Yagi antennas usually have self-supporting elements. These observations led me to expect that a cleverly designed Yagi antenna was the way to proceed.

construction details

This antenna design represents a compromise between locally available materials, package size, and antenna performance. I decided to build a four-element Yagi which is assembled something like a custom Erector™ Set. The boom and mast are each made from pieces of aluminum angle-stock. This allows the pieces to nest together when the antenna is packed. The elements are made of pieces of small diameter aluminum tubing. By making the individual boom pieces 16 inches (406 mm) long, three pieces can make a 48-inch (1220-mm) boom. This is a reasonable size for a four-element Yagi on 2 meters. Also, by making the element spacings 16 inches (40 cm), the centers of the driven element and first director will be at joints in the boom, leaving fewer places where parts have to be joined.

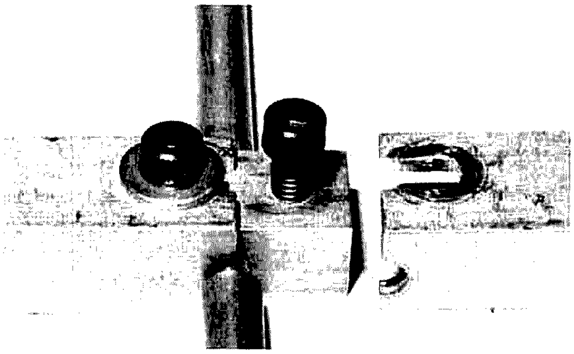
Having established the element spacings and diameters for mechanical reasons, I next needed to calculate the element lengths. Fortunately, I have a computer program for Yagi antennas. It includes an optimizer routine, which allows the computer to systematically try many combinations of antenna dimensions to find those that would give good performance.

table 2. Possible gain for Yagis and quads with different numbers of elements.

number of elements	Yagi gain		quad gain
1	0 dB	(reference)	2 dB
2	5-6 dB		7-8 dB
3	8-9 dB		10-11 dB
4	10-11 dB		12-13 dB
5	11-12 dB		13-14 dB

For this particular antenna, I was most interested in achieving gain over the entire 2-meter band. Front-to-back ratio was not considered important. There are many combinations of element spacings and lengths which could be expected to give similar performance. However, the spacings were chosen for mechanical reasons. Furthermore, by making the two directors identical in length, the possibility for errors when assembling the antenna is eliminated. These compromises probably cost a dB or so over an antenna intended for maximum gain at one frequency, but were considered worthwhile.

As mentioned, aluminum angle-stock is used for the boom and mast. The boom is made from three pieces of 1/2 × 1/2-inch (13 × 13-mm) angle. Each piece is 16 inches (406 mm) long. Hence, the assembled boom is 48 inches (1220 mm) long. Similarly the mast is made from three pieces of 1/2 × 1-1/2-inch (13 × 38-mm) angle. The pieces of the mast and boom are joined by small aluminum blocks and 8-32



Close-up of the joint between the first director and the boom. The boom is made from lengths of aluminum angle-stock which are fastened to the block by screws. The element halves screw into the sides of the same block.

The boom is assembled and the first part of the mast is added. (Photo courtesy N6ST.)

cap screws. Slots in the ends of the aluminum angle allow the pieces to slide apart when the screws are loosened, without being completely removed. Keeping the screws in the blocks reduces the effort needed to reassemble the antenna. The cap screws can be hand-tightened adequately for temporary use. I carry a small hex-wrench to tighten them more securely for longer operating periods.

The elements are made of 1/4-inch (6-mm) aluminum tubing. The eight tip sections are each 16 inches (406 mm) long. The center sections are 2 inches (50.1 mm) long for the directors, 3 inches (76.2 mm) for the driven element, and 4 inches (101.6 mm) for the reflector. Making the center sections different lengths makes it very easy to put them in the correct place on the boom. The correct tip section is always the top piece on the pile.

On each of the sixteen element-pieces, the end towards the boom has a permanently attached 8-32 thread. This was done by first tapping a screw thread inside the tubing. Next the end of a 0.5-inch (12-mm) headless set-screw was dipped in epoxy. Then the set screw was threaded into the end of the element piece until about 0.25 inch (6 mm) was exposed. After the epoxy set, the screw was permanently fixed.

The outer end of each of the center sections has an internal 8-32 thread to receive the screw from the tip section. This thread is installed by reaming the inside of the tubing to the correct diameter and putting a steel-threaded insert in the tube. These inserts are commonly sold to repair threads which have been stripped. Here the insert protects the aluminum from wear as the antenna is assembled and disassembled. The thread size was chosen to be compatible with the tubing-wall thickness and inside diameter. You may well find that a slightly larger or smaller size is better suited to your tubing.

Assembly is begun by lining up the boom pieces and tightening the screws. Then the mast is assembled and connected to the boom. Next, the center sections of the elements are screwed into the sides of the same blocks which join the boom pieces. Finally, the element tips are put in place.

The antenna can be easily assembled or disassembled in under five minutes. At current prices, all of the material costs about $10 at the local metal supplier. The whole thing weighs under two pounds, which is certainly less than an amplifier and power supply.

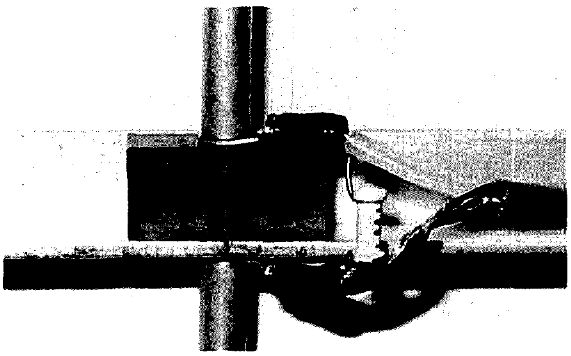
table 3. Final element lengths.

Director 1	36.5 inches	927 mm
Director 2	36.5 inches	927 mm
Driven Element	38.5 inches	978 mm
Reflector	40.5 inches	1029 mm

feedline matching

Nothing has yet been said about connecting the feedline. The center of the driven element is a Plexiglas™ block instead of the aluminum blocks used elsewhere. The driven element is fed as a center-fed dipole. With no matching circuit, the SWR is about 6:1.

The original plan was to make a small circuit board with a suitable impedance-matching circuit. This



Close-up of the driven element and matching circuit. The mechanics are the same as Photo 2 except that the block is acrylic plastic and the coax cable is connected to the two sides of the driven element. The tubular capacitor is used for impedance matching.

approach was expected to be smaller and lighter than a gamma match or some of the other impedance-matching methods. After some experimenting with different matching circuits, it turned out that a simple capacitor is all that is needed. Without matching the driven element would present an impedance consisting of a small resistance and inductance. The addition of a 10-pF capacitor across the antenna terminals provides a VSWR of 1.5 to 1 to the feedline (over the entire band).

Experiments were made with the center conductor of the coax connected to the top side and the bottom side of the driven element. The antenna seemed to work better with the top side connected to the center conductor. Possibly there is some interaction with the metal mast which is on the bottom of the antenna. Experiments were also made with and without a balun. The balun does not seem to offer any improvement, and so is not included in the final design.

The author carrying the complete station to a hilltop operating site. (Photo courtesy N6ST.)

performance measurements

Antenna gain was checked in two ways. The first way was to switch between a dipole and the beam while asking the receiving station for a comparison. This yielded reports as high as 20 dB.

More reliable measurements can be made comparing the received signal strength. A switchable attenuator should be put in the feedline. A moderately strong signal is then tuned in, and the attenuator adjusted until the signal just breaks the receiver squelch. Next, the beam antenna should be connected and the attenuator readjusted until the signal breaks the squelch. The difference (in attenuator readings) is antenna gain. For tests in clear locations, the gain measures about 10 dB, as expected.

Under conditions of multi-path propagation, results are less consistent. Small changes in the position of the reference dipole make a big difference in the received signal-strength.

However, since multi-path propagation is a common occurrence on 2 meters, let's consider it for a moment. In multi-path propagation, obstacles and reflecting objects cause the signal to reach the receiving antenna from two or more different directions. For simplicity, consider the extreme case where there are two signals of equal strength. At some antenna locations, the signals are out of phase and cancel. In this case no net signal will be picked up by the antenna. At other locations the signals will be in-phase and add. The antenna will pick up a total signal which is 6 dB stronger than if the antenna only picked up one of the signals. Under these conditions, a carefully placed vertical dipole could equal the performance of a directional antenna with 6-dB gain.

This is the type of effect which I have observed with the portable beam. Under conditions of severe multi-path propagation, it does not have the 10-dB gain over a dipole. However, the beam does have a different advantage: it is much less sensitive to position than the dipole. In trying to raise a distant repeater, aiming the beam in the right direction and making one transmission is all that is necessary. With the dipole, several attempts may be needed to find a good spot. Even then, a 10-dB beam still has an advantage in signal strength.

alternative construction ideas

I would like to suggest two other ways to build a portable beam. First, instead of tubing the elements could be made from pieces of metal measuring-tape. The tape would be strong enough to hold itself up when the antenna is in use. The elements could then be coiled up for carrying.

A more exotic scheme would be to build the antenna on a sheet of Mylar plastic with elements made of strips of aluminum foil. Such an antenna could be folded up and put in your shirt pocket. The difficulty with this design is finding a way to hold the antenna up when you wish to operate, and keeping it from blowing away in a breeze.

conclusion

I am sure that any of these three ideas will make your 2-meter portable operations more enjoyable.

ham radio

achieving the perfect VHF antenna null

Principles borrowed from a-m broadcasters permit steerable nulls with theoretically infinite attenuation.

With fixed-location VHF stations, such as repeaters, a situation sometimes occurs where more than one station is received on a given channel, and one of them must be rejected. A common solution has been to use a directional antenna such as a Yagi. However, a single antenna may not provide the required signal rejection.

For years, standard a-m broadcast stations have used directional antenna systems to solve interference problems. The principles involved are applicable not only to the standard a-m broadcast band, but also to VHF antenna systems. Many problems can and have been solved using only two antennas.[1,2,3,4]

design considerations

Several factors are important in the design of an antenna system capable of peaking signals from one direction while nulling those from another. For peaking, two signals must be in phase. For signal nulling, the basic requirement is having two signals that are equal in amplitude and have a phase difference of 180 degrees.*

*Theoretically, *any* number of signals can add up to a zero amplitude result (signal). However, it rapidly becomes more difficult to null increasing numbers of independent, time-varying signals. **Editor.**

A two-antenna system that provides a peak in one direction and a null in another is shown in **fig. 1.** Signals from direction A arrive at both dipoles (horizontal or vertical) at the same time. The spacing between the two antennas cause signals from direction B to arrive at antenna 2 with a time difference equal to one-half wavelength, equivalent to a 180-degree phase shift. If both antennas are fed in phase (equal length feedlines), signals from direction A add while those from B cancel.

The equation for determining required spacing is:

$$S = \frac{5904}{f\,|\sin\alpha|}$$

where α is the angle between the desired signal and the undesired signal directions; f is the frequency in MHz; and S is the antenna separation in inches.

The nulling arrangement works well with practically any antenna — horizontal dipole, vertical dipole, and Yagi, etc. The angular displacement between the directions of the two signal sources may be anything from 0 to 360 degrees. Since the same absolute value of the sine function occurs four times over a complete rotation, any pattern is symmetrical and exhibits four separate nulls. Consequently, the required spacing for an angular displacement between signal sources of 45 degrees is the same as that required for one of 135 degrees, 225 degrees, or 315 degrees. Other antenna separations, such as odd multiples of S, can provide the same results. However, there is a limit to practical applications of this system. The spacing required for angular displacements around 0 degrees and 180 degrees becomes too large to implement.

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more practical nulling methods

Required mechanical tolerances for antenna placement can be relaxed if additional techniques, such as electronic control of phase-shift and amplitude, are employed. The exact 180-degree phase shift for the undesired signal may be set by feedline length, and variable gain preamplifiers may be used to provide two signals of equal amplitude. An adjustable feedline design[5] that provides a continuous phase shift is illustrated in **fig. 2**.

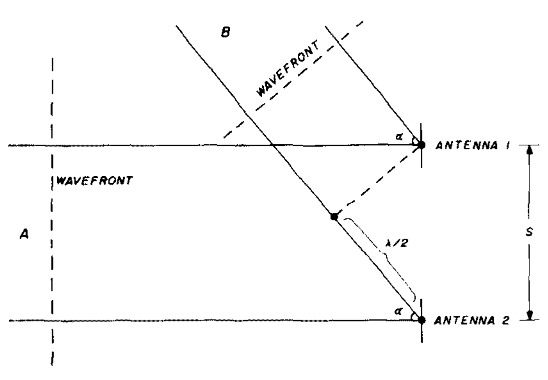

fig. 1. Example of simultaneous peaking (from direction A) and nulling (from direction B) using a two-element array.

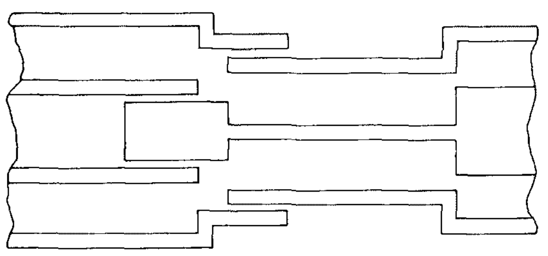

fig. 2. Adjustable-length coaxial line section with constant impedance.

Construction of this is not a simple task. However, a small amount of error is tolerable; **fig. 3** is indicative of a practical feedline design. Many hobby stores stock, or can obtain, brass tubing with a wall thickness of 1/64 inch and diameters at 1/32-inch gradations. Adjacent sections telescope together, and by proper selection of tubing size for the inner and outer conductors, the characteristic impedance of each section can be set to approximate either 50 ohms or 75 ohms at a unity velocity factor. As it works out, type F and BNC connectors are well-suited for mounting into the ends of these. For some of the smaller diameter units, however, it is necessary to file down the end of the connector for best fit.

To assure a solid assembly, the flange of each connector should be spot-soldered to the tubing. A string is clamped to each end, slightly shorter than the maximum extended length of the section, to prevent the section from separating into two pieces during adjustment. **Table 1** lists practical combinations of tubing for use with type F and BNC connectors.

Amplitude match, the second condition, is obtained using the preamplifier shown in **fig. 4**. The preamplifier uses an untuned input circuit to reduce gain variations prior to signal combining. Any preamplifier instability can be reduced by placing a low-value resistor (10-27 ohms), or ferrite bead, in the drain lead of each J310.

A complete system that uses Yagi antennas in the array appears in **fig. 5**. The phase section and preamplifier unit were adjusted using signals in the fm broadcast band. In many cases signals could be null-

table 1. Brass tubing combinations for practical adjustable-length sections using 75-ohm and 50-ohm coaxial cable.

conductor	section A	impedance (ohms)	section B	impedance (ohms)	average impedance (ohms)
outer	13/32 OD	65.9	12/32 OD	77.9	71.9
inner	4/32 OD		3/32 OD		
outer	14/32 OD	70.6	13/32 OD	83.1	76.9
inner	4/32 OD		3/32 OD		
outer	10/32 OD	65.8	9/32 OD	83.1	74.4
inner	3/32 OD		2/32 OD		
outer	14/32 OD	46.3	13/32 OD	52.4	49.3
inner	6/32 OD		5/32 OD		

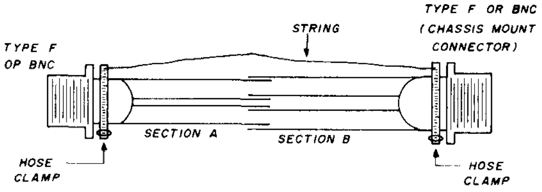

fig. 3. Adjustable-length coaxial line section using brass tubing with each half approximating impedance of associated cable.

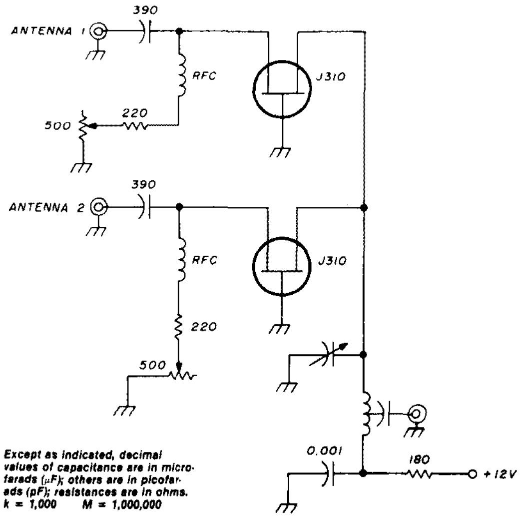

fig. 4. Diagram of preamplifier unit for adjusting signal level from each antenna.

ed down to noise level. In some cases nulling one station revealed another station on the same channel.

system limitations

The system is not totally effective in cases of multipath, where the undesired signal arrives from more than one direction. Nor is the system totally effective if the antennas are not rigidly mounted. It takes very little physical displacement to upset a perfect null setting; this means antenna rotators cannot be used as they provide too much variability in setting, as well as backlash.

Although the antennas are electronically finetuned for nulling the undesired signal, such care, as mentioned before, is not required for peaking. An error of as much as 10 electrical degrees from boresight reduces the gain by only about 0.3 dB.

special case: 180-degree displacement

If the signal to be rejected is coming from the direction opposite the desired signal, another configuration can be used. **Fig. 6** shows two antennas, a quarter-wavelength apart and fed 90 degrees out-of-phase. Signals from A hit antenna one 90 electrical degrees before they hit antenna two. Since the feedline from antenna one is 90 electrical degrees longer than that from antenna two, signals from direction A arrive in-phase. On the other hand, signals from B hit antenna one 90 electrical degrees after they hit antenna two. They are further delayed another 90 degrees by the long feedline to antenna one, giving a total phase shift of 180 degrees. Spacings at any odd multiple of a quarter-wave also provide nulling.

Fig. 7 shows how Yagis may be used in a system exhibiting infinite front-to-back ratio. Again, the peaking criteria need only be approximated, while the system is electronically fine-tuned to give total nulling of the signal off the back. Slight shifts in spacing and/or feedline length may be used to generate nulls in the vicinity of 180 degrees. The null can be slewed off the 180-degree direction by changing

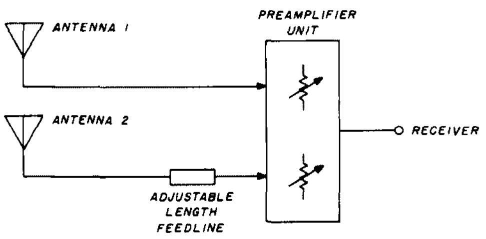

fig. 5. Practical antenna system with electronic null adjustment.

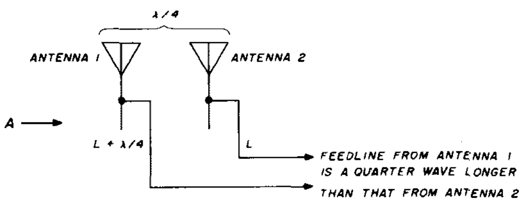

fig. 6. Dipole pair setting for peaking signals from one direction and nulling those displaced 180 compass degrees.

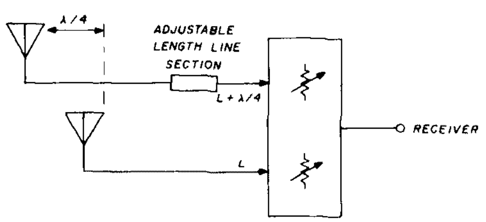

fig. 7. Practical antenna system with "infinite" front-to-back ratio.

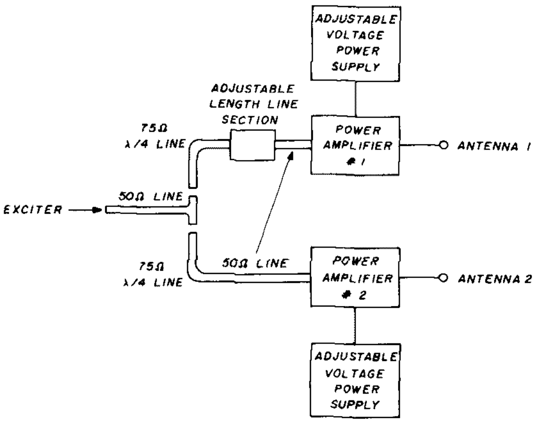

fig. 8. Null adjustment with a transmitting system.

antenna spacing, feedline length, or both. Another method is to maintain the 180-degree null point and to aim the back of the array toward the undesired signal.

the first system, or at some odd multiples of a quarter-wave in the second system, has value. The closer together two antennas are placed, the more they interact. Wide spacing effectively reduces this interaction.

transmitting arrays

Null principles can be applied to transmitting systems as well. **Fig. 8** shows how two power amplifiers and an adjustable-length line section can be used to secure a perfect null in a transmission pattern. Feedline length from each amplifier to the antenna should be equal. Amplifier input lines, as measured from their common junction, should be equal in length, or have a difference of one-quarter wavelength, depending on the system used.

temperature changes influence pattern

Another factor to consider is the effect of temperature on feedline length. This has been a problem in broadcast applications.[6] The effect may be minimized by making the outdoor portion of each feedline section equal in length. The adjustable-length line section and preamplifier unit are best located indoors near the receiver, protected from the elements. Here they can also be adjusted by observing a local field-strength meter. There should be minimal signal pick-up by the feedlines, as any direct signal pick-up by a feedline serves to mask the true antenna pattern. It has been reported that military RG cable provides about 35 dB shielding, whereas less expensive cable may provide only 20 dB. Full braid, duofoil, or double-shielded coaxial feedline may be necessary in difficult situations.[7]

Finally, the nulling criteria holds only for a single frequency. However, attenuation remains high around the set frequency. For example, if the null is set for the carrier of an fm broadcast station, which has a channel of + / − 100 kHz, the calculated attenuation decreases from infinity at the carrier frequency to about 80 dB at the channel limits.

The systems described here can solve many problems. If there is another signal on a desired repeater's frequency, it can be nulled out. Setting a null in a transmitting pattern may offer a solution to an rfi problem. Then too, the systems may be used to reduce interference in fm broadcast or TV station reception. An interesting application is nulling one of the desired signals in a multipath distortion problem. By adding two coaxial relays it is possible to expand the system to the capability of switching the null from one direction to another. For example, the addition of a half-wave section in either feedline may be used to reverse null and peak directions. The adjustable-length line section and preamplifier are also effective with circularly polarized antenna systems, since they permit total nulling of signals of one sense or the other.

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ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

Antenna experimentation is one of the few fields in which an Amateur can participate armed only with enthusiasm, a tape measure, an SWR meter and inexpensive tools. No Ph.D. degree in higher mathematics or computer technology is required.

One of the best candidates for home experimentation is the quad antenna (**fig. 1**). The quad loop can be built in many configurations. The support structure can be as uncomplicated as a set of bamboo poles and the whole arrangement can be built for only a few dollars. A single loop parasitic element added to the driven loop makes a two-element quad beam. In many areas of the world where aluminum tubing is hard to find, or prohibitively expensive, the quad antenna is the best answer to the need for a high-gain, high-frequency antenna.

the single-element loop antenna

While the loop antenna has been known since the early days of radio, the use of a large loop for hf transmission was not seriously investigated until 1938 when Clarence Moore, ex-W9LZX, developed a two-element loop antenna for shortwave broadcasting. The Moore design was an instant success and the so-called quad antenna has been popular with Amateurs worldwide for the past four decades.

The simplest quad is a single loop which provides horizontal polarization when fed as shown in **fig. 1**. The loop has a bi-directional pattern similar to that of the dipole. Loop gain and feedpoint impedance are a function of the shape of the loop. The loop having the highest gain and feedpoint resistance is the circular model. This provides a power gain of about 1.13 dB over a dipole with a feedpoint impedance of 135 ohms. The square design has a gain of about 0.85 dB over a dipole and a feedpoint impedance of 120 ohms. The triangular, or "delta," loop provides a gain of about 0.55 dB over a dipole and a feedpoint impedance of 105 ohms.

An intermediate-design loop which provides a power gain of 1.5 dB over a dipole and a feedpoint impedance of 50 ohms is shown in **fig. 2**. This quad loop (while a bit unwieldy for the lower frequencies) is an excellent antenna for the higher bands, as it provides bi-directional gain and can be fed directly with a 50-ohm coaxial line. A similar design, to match a 75-ohm line, is also shown.

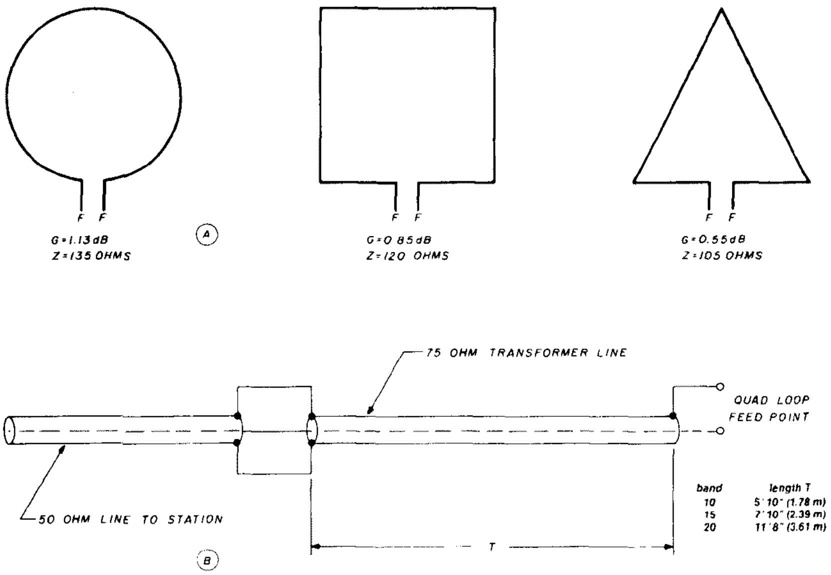
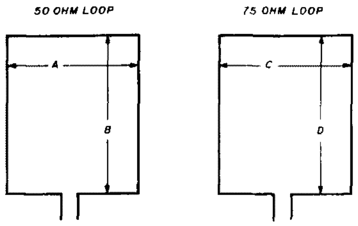


fig. 1. *(A)* The simple transmitting loop. Directivity is in and out of page. The triangular loop may be inverted, with apex at bottom and feedpoint at apex. *(B)* Quarter-wave transformer for use with quad loop antennas.



band	A	B	band	C	D
10	5'6" (1.7 m)	11'8" (3.6 m)	10	6'10" (2.1 m)	10'4" (3.1 m)
15	7'5" (2.3 m)	15'9" (4.8 m)	15	9'3" (2.8 m)	13'11" (4.2 m)
20	11'1" (3.4 m)	23'7" (7.2 m)	20	13'10½" (4.2 m)	20'10" (6.3 m)

fig. 2. Single element quad loops for 50- and 75-ohm feedlines.

The delta loop and the circular loops have a feedpoint impedance somewhat different from that of the square, but all of these designs can be nicely matched to a 50-ohm transmission line by the use of a quarter-wavelength, 75-ohm transformer between the line and the loop. Data for such a transformer is given in **fig. 1**.

The loop antenna is balanced to ground at the feedpoint and it is a good idea to isolate the outer shield of the coaxial feedline from antenna current. This can easily be done by winding the line into a four-turn coil about 8 inches in diameter directly below the loop. The plane of the coil should be at right angles to the plane of the loop.

One of the advantages of the loop antenna is that it can be supported at the midpoint by a single pole. Properly built, the loop is not obtrusive and can be used in areas where more conspicuous ham antennas are frowned upon.

The 50-ohm or 75-ohm loop can be turned on a side to provide a vertically polarized array for low-frequency operation. For 40 meters, for example, loop height is only about 22 feet, and the extensive radial system that is required for a ground plane antenna is not as necessary (see **fig. 3**).

the cubical quad beam antenna

Adding a parasitic element to the driven loop produces the famous cubical quad antenna pioneered by ex-W9LZX. The quad is a unidirectional array providing a power gain of about 6 to 7 dB with a good front-to-back ratio. Both gain and f/b ratio depend upon element separation and tuning, as is the case with the traditional Yagi beam design.

It is difficult to surpass the advantages offered by the simple two-element quad. It is light and has low wind resistance, and it provides high gain in a small package. The feed system is uncomplicated. In addition, since the elements are continuous and have no tips, rain static problems (often a headache with the Yagi beam) are nonexistent. The cubical quad beam is thus an ideal antenna for the DXer who wants to get good results with a minimum expenditure of money.

a practical two-element cubical quad

Data for a practical two-element quad are given in **fig. 4**. Boom length is about 0.12 wavelength, which provides a compact design and a good match to the coaxial transmission line, since feedpoint impedance of the quad is a function of element separation as well as tuning. The reflector loop is pre-cut to the correct dimension and requires no adjustment after assembly. Important dimensions are shown in the illustration, the length **R** being the distance from the center point of the assembly to the point of attachment of the wire to the support structure.

The crossarms for the quad should be made of insulating material. Many quad assemblers have run into problems when metal arms are used for the array. It is possible to insert insulating sections in metal crossarms, but the builder is advised to stay away from this complicated technique. Fiber glass poles, bamboo, and PVC pipe have been used successfully for quad arms.

Most homemade quads use a section of 2- or 3-inch diameter aluminum tubing for the boom. The two-element quad usually requires 2-inch tubing, but a quad for 6 or 10 meters can use a smaller diameter boom.

Boom-to-crossarm clamps are available from several manufacturers, but many builders have made their own out of a plywood sheet and galvanized-iron angle brackets. If you take this approach, make sure that the edges of the plywood are sealed against moisture penetration. Two or three coats of outdoor house paint will do the job.

A more exotic design makes use of a "spider" arrangement which employs multiple crossarms supported from a central point on the mast, at the middle of the array.

how high the quad?

Experience has proven that the quad antenna will perform well even though mounted close to the earth. As an example, the main lobe of a quad antenna mounted one-quarter

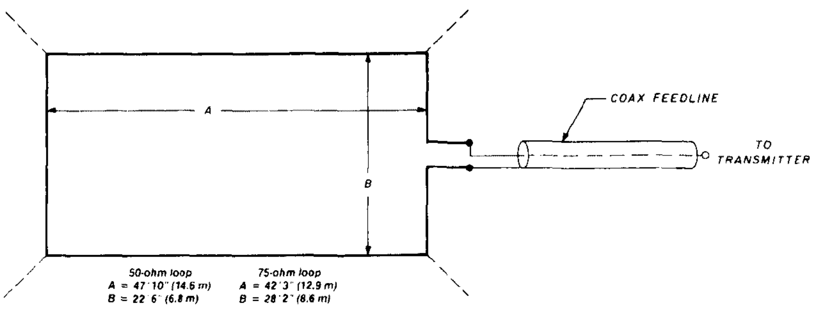
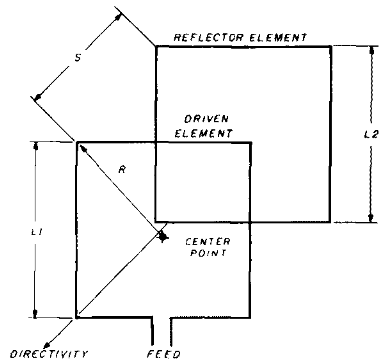


fig. 3. Vertically polarized 40-meter loop for 50- or 75-ohm feed. Mount loop in vertical plane as high above ground as possible. Bring feedline off horizontally.



band	side dimension $L_1 = \frac{250}{f(MHz)}$	side dimension $L_2 = \frac{258}{f(MHz)}$	spacing $S = \frac{118}{f(MHz)}$	dimension R
40	35′2″ (10.72 m)	36′4″ (11.07 m)	17′0″ (5.18 m)	24′8″ (7.52 m)
20	17′8″ (5.38 m)	18′2″ (5.54 m)	8′5″ (2.56 m)	12′6″ (3.81 m)
15	11′8″ (3.56 m)	12′3″ (3.73 m)	5′7″ (1.70 m)	6′4″ (1.93 m)
10	8′8″ (2.64 m)	9′1″ (2.77 m)	4′2″ (1.27 m)	6′2″ (1.88 m)

fig. 4. Design data for two-element quad. Dimension R is approximate distance from center point of loop assembly to point of attachment of wire.

wavelength above the ground is at an elevation angle of 40 degrees, whereas the angle of maximum radiation of a dipole at the same height is straight up.

At a height of three-eighths wavelength the angle of radiation of a quad is about 32 degrees below that of a Yagi or dipole at the same height. Finally, at a height of one-half wavelength, the radiation angle of the quad and the dipole (or Yagi) are about equal. (The height of the quad is measured to the bottom of the lower element, as that is the point at which the quad is usually supported).

The upshot of this is that the quad does better in terms of low elevation angles than does either the dipole or the Yagi beam. True, a height of one-quarter wavelength is not a good one as far as low-angle, long-distance DX is concerned, but if you are stuck with it, it is better to use a quad than almost any other antenna because of the lower angle of radiation.

Those Amateurs lucky enough to get the quad up in the air from 40 to 60 feet above ground will quickly find out why the quad achieved world-wide popularity in a very short time. Build a quad and enjoy!

RFI revisited — 18 MHz

The 18-MHz band (18.068-18.168 MHz) has not been opened for general use in the United States, although Amateurs in several other countries are already using it on a non-interference basis. Use of the band in the U.S. poses some interesting problems so far as RFI goes. The third harmonic of the band (54.2-54.5 MHz) falls extremely close to the video (picture carrier) frequency of television channel 2 (55.25 MHz).

This situation is unique; I can't think of another circumstance where the harmonic frequency of an Amateur band falls so close to a television video channel.

My experimental license (KM2-XDW) permits restricted operation in the 18-MHz band, and this provided the incentive to explore the question of TVI on this new ham band. One of the first experiments I ran on 18 MHz was to determine the degree of TVI that I would encounter when operating on this band. I used my regular station equipment, which included TVI suppression techniques such as a lowpass filter in the transmission line, bypassed power lines, and good equipment grounding. This sufficed to provide adequate TVI protection on all Amateur bands when the TV receiver was equipped with a highpass filter. Alas, operation on 18 MHz quickly pointed out that ordinary TVI suppression was insufficient in my case to reduce channel 2 television interference to an acceptable level. After a few false starts, however, I was able to clean up the problem, which seemed to be a combination of fundamental overload plus harmonic interference. Here's what I did:

First: I wound about five turns of the transmission line (RG-58/U) at the transmitter around an iron-powder toroid core of 2¼ inch diameter (Amidon T-225-2). This was done to "cool off" the outside of the coaxial line to the antenna. A similar toroid choke was placed at the antenna end of the line.

Second: The garden variety highpass filter on the television set was replaced with a higher attenuation unit (*J.W. Miller C-513-T3* for 300-ohm line, or *C-513-T2* for 75-ohm coaxial line). These filters provide about 60 dB of attenuation to signals below 40 MHz.

Third: The line cord of the television receiver was wrapped around a ferrite core, similar to the one used on the transmitter feedline. This was done to isolate the receiver from rf picked up by the power line.

After these three fixes were incorporated into the station, the television receiver was reasonably clear during 18-MHz operation, even at a kilowatt input level. I was transmitting into an antenna only about 18 feet away from the TV antenna.

It was interesting to note that some TVI measures actually degraded the TV picture. One brand of TV filter, for example, when placed in the ribbon line, seemed to upset the TV tuner, as it produced "sound bars" on the picture which wiggled about with the audio signal. Removing the TVI filter and replacing it with the one specified cleaned up the wiggly lines.

Grounding the TV receiver chassis (through a 0.01-μF, 1.6-kV disc capacitor for protection) increased the TVI level, possibly because the ground lead was long enough to act as an antenna at 18 MHz.

In summary, it is possible to clean up TVI at 18 MHz, but it takes special care to make sure the transmitter is "clean" for channel 2 reception. In addition, the television receiver has to have a good highpass filter in front of it to provide maximum overload protection from the transmitter.

references

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2. For additional information on TVI and RFI, read, "Interference Handbook," available for $8.95 plus $1.00 shipping from Ham Radio's Bookstore, Greenville, New Hampshire 03048.

ham radio

An effective DX antenna
that's easy to put up
— and that stays up

four-vertical collinear element 20-meter array

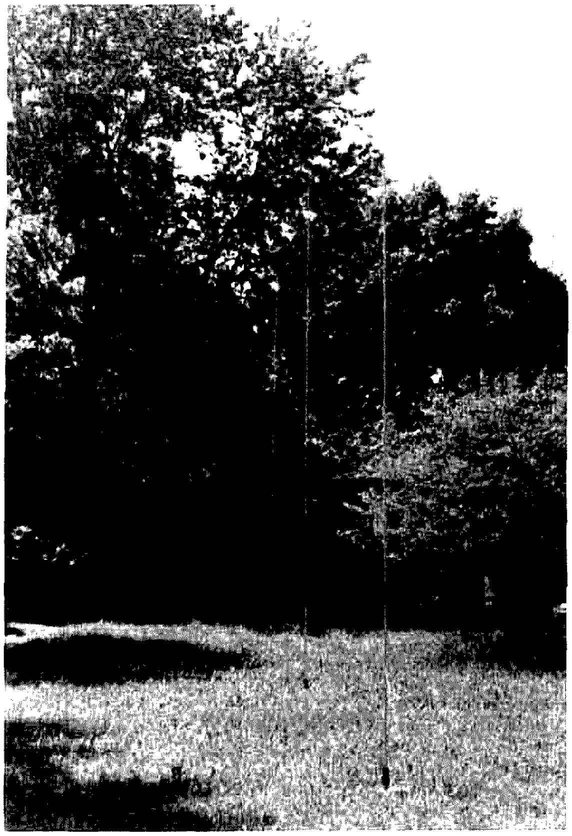

fig. 1. 20-meter phased array.

This is a 20-meter version of the 80-meter array described in *QST* in 1965. It represents one method of providing directional performance without the use of a rotator. Interconnect figure courtesy ARRL. Editor

I had never been impressed with vertical antennas until I phased a pair of 40-meter quarter-wave verticals a few years ago. Since the two worked so well, it seemed reasonable that four should work even better. I constructed a phasing box for four in-line vertical antennas.[1] However, *not having the time* to erect this system, I stored the relay box away.

A *job change* some time later brought me to a small ranch duplex adjacent to an open field. I erected a single 20-meter quarter-wave vertical in the middle of the field using a ground system consisting of eight 16-foot-long three-conductor radials.

four-element array construction begins

Soon after this I started gathering parts for the four 20-meter verticals. Using pieces of 1-inch (25.4-mm), 7/8-inch (22.23-mm), and 3/4-inch (19.05-mm) aluminum tubing with 0.058-inch (1.45-mm) walls, I constructed four 16-foot 6-inch radiators using stainless steel automotive hose clamps and a slit tubing

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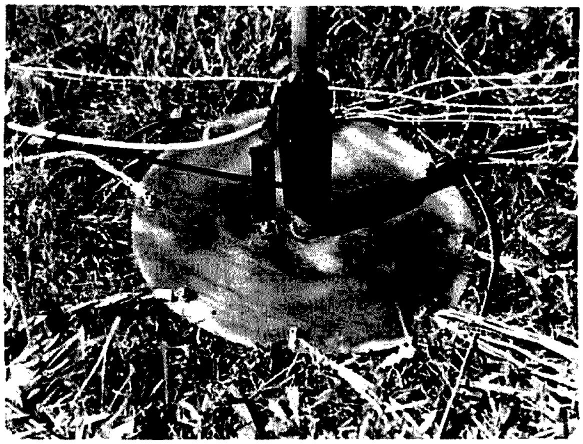

fig. 2. Basic ground system under each radiator, consisting of copper disk and No. 16 insulated ac house wire.

technique. The base insulators were old plastic spacers from a 20-meter quad. The antenna was mounted on 1-inch treated-wood dowels driven several feet into the ground (fig. 1).

The ground buss consists of surplus copper disks from a junk yard. A 1-1/8-inch (28.56-mm) hole was cut in the center of the disk and a series of holes drilled around the perimeter with radials attached to them by brass nuts and bolts (fig. 2). The radials were number 16 insulated ac house wire. Finally, each disk, as well as the antenna connections was given two coats of clear Krylon® to retard corrosion after radial wires were attached.

The verticals were laid out in line from northwest to southeast, the switchable end-fire directions. When the two broadside lobes were switched in, two squashed figure-eight lobes resulted, one on southern Europe and the other on the South Pacific. Since I was mostly interested in working into Asia, I considered this the best compromise.

The verticals were spaced 16-feet 6-inches (5.03 m) apart and each was fed by equal three-quarter wavelength RG-8X coaxial lines. The main feeder, power divider, and three phasing lines used RG-8. The ground systems consisted of four single-conductor quarter-wavelength wires under each antenna, making it difficult to work into Asia. The small ground system adversely affected the array performance. After adding eight three-conductor 16-foot 6-inch (5.03-m) radials to the original four wires, (a total of twelve radials) I noticed 4 to 6 dB difference in transmission and a bit better front-to-back ratio on receive. Knowing the importance of a good ground system and with a future 40-meter installation in mind, I laid an additional thirty 33-foot-long radials under the two outer (NW) verticals, in about the 120-degree sector. A total of forty-two radials were now connected to the outer antennas.

The VSWR using only twelve radials was NW - 1.2:1; SE - 1.4:1; broadside - 2.4:1. With the addition of thirty 33-foot-long radials under the two outer antennas, the VSWR was reduced to NW - 1.05; SE - 1.15:1; broadside - 2.01:1.

The relay phasing box, fig. 3, is wired as shown in fig. 4. Internal leads should be kept as short as possible. When constructing the relay lines, phasing harnesses, and power dividers, remember that the velocity factor of coax can be 0.66, 0.77, and sometimes 0.81. It pays to check what the VF is before you start cutting the coax. The electrical length of the phasing lines is $\frac{246 \times VF}{freq.\ in\ MHz}$ for a 90-degree or one-quarter wavelength line.* For the 180-degree or 270-degree lines, just multiply by a factor of two and three respectively. I used type-N connectors and a type-N female T-connector for the power divider since they are waterproof and constant impedance devices. I found the rubber boots for the phasing box connectors at a hamfest. The RG-8 coax and relay wire (inexpensive doorbell wire) was placed along a neighbor's fence. I used surplus 50-cycle 120-Vac large-contact relays that actuate at 35 Vdc.

The vertical array is easy to access (phasing box and antenna connections) and maintain. If a 16-foot radiator falls down as a result of heavy winds or ice loading, it can be rebuilt easily.

performance

I worked two VK stations, both running little Heathkit HW-8 QRP transceivers! On checks with UAØWAY and UA9OH running just the 100-watt

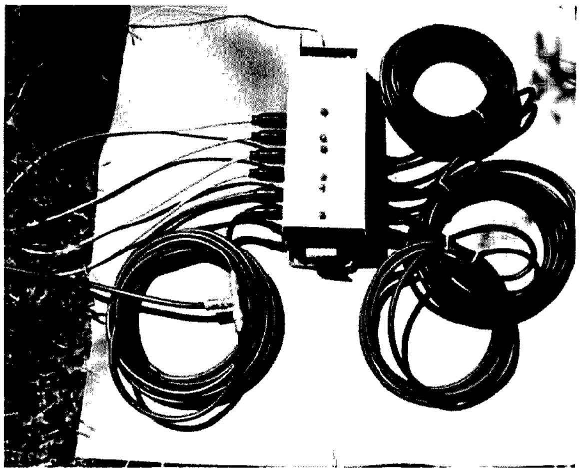

fig. 3. Relay box, phasing lines, and antenna input power divider (T-connector).

*See assumption 5 on page 20.

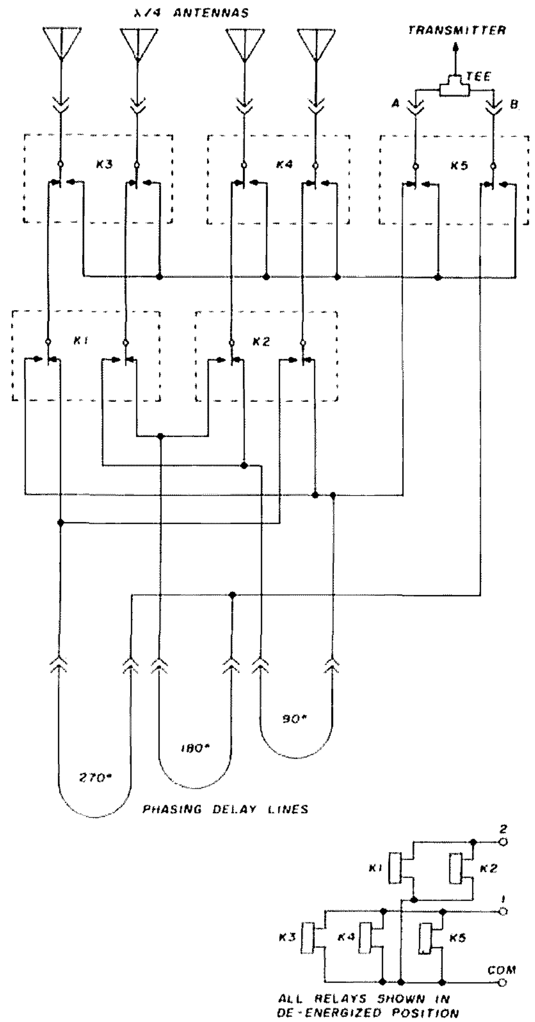

fig. 4. Relay inconnection diagram for the four-element phased array. The preferred end-fire direction is present with no dc applied. Voltage applied to terminals *2* and *COM* reverses the array (still end-fire), while dc applied to terminals *1* and *COM* provides a bi-directional broadside lobe pattern.

transceiver on SSB, front-to-back was in excess of 30 dB and sometimes as high as 40 dB. This is helpful when you're trying to reject southern QRM and looking for a weak 9V1 or 9M2 station over the North Pole.

reference

1. Dana W. Atchley, Jr., W1WKK, "A Switchable Four-Element 80-Meter Phased Array," *QST*, March, 1965.

ham radio

inexpensive connectors for hardline

Hams combine ingenuity and plumbing fittings to solve costly interface problems

A great deal of surplus hardline has recently become available from CATV companies at very low cost. The hardline has a solid aluminum outer shield with either a solid copper or copper-clad aluminum center-conductor. This high quality, low loss, VHF/UHF cable is great for repeater or home stations. There is only one problem — connectors are expensive, *if* they can be found. Once again, ham ingenuity and homebrew construction are necessary.

I needed a connector (for 1-inch cable) which would be simple and cheap to manufacture. Designing one required some thought and many hours' rummaging through local plumbing suppliers' stock. It takes only about 10 minutes to make each connector. The cost per connector is about $2.00 — far less than they could be bought new. Construction is not hard, and you may use considerable latitude choosing materials.

First check out your local plumbing stores to see what is available. The fittings I used were (1) a 3/8-inch threaded to 3/16-inch tubing (nipple) adapter (this may be called a barb); (2) a 3/4-inch threaded female to 1/2-inch copper tubing adapter; (3) an SO-239 coaxial connector. These are shown in **fig. 1**, along with a section of the 1-inch line.

construction

Some machining is required to make the center of the adapter. I have a Shopsmith Mark V that I used as a lathe. It is possible to do the same thing using a

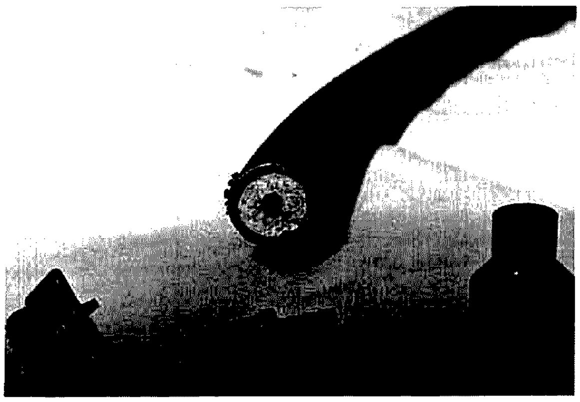

fig. 1. The 1-inch hardline, coax connector, and the plumbing fittings used to make a connector for the hardline.

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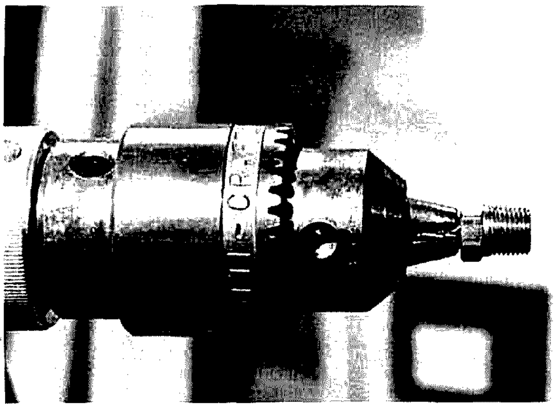

fig. 2. The Barb fitting prior to machining.

standard 1/4-inch drill mounted in a vise or stand. There is no high torque or stress involved, so either method is fine.

The first step is to chuck up the nipple adapter with the nipple end in the chuck. Make sure it is centered in the chuck! This step is shown in **fig. 2.** Start the lathe (drill) at a moderate speed. First, using a coarse and then medium file, machine away the flat surfaces. Then file down the threaded section. After a single cylinder is obtained, use a fine file to smooth the assembly. The final outside diameter should be 5/16-inch (7.94 mm). Then, very carefully, use a rat-tail file to taper out the inside of the fitting. The reason for this taper is to ensure a good press fit against the center conductor when the completed connector is placed on the line.

Now stop the lathe and reverse the fitting in the chuck. **Fig.** 3 shows this step. You can see how the large end has been machined. Again, the adapter must be placed squarely in the chuck. Using a medium and then a fine file, smooth out this piece and round off the shoulder slightly.

The next step requires some dexterity. A small vise and some clamps help. Fit the small end of the machined adapter into or over (depending upon the exact fitting and connector you use) the center connection of the SO-239. Solder the two pieces together, making sure the fitting fits squarely on the SO-239 (**fig. 4**).

Now use some fine sandpaper to clean the small end of the large reducing-fitting and the SO-239. Apply a small amount of soldering flux to the SO-239 body and the large reducer. Remember that these are plumbing fittings and not wires you're soldering; if you omit this step you'll find out why plumbers always use flux. Press the SO-239 into the adapter. This should be a close fit, requiring only hand force to assemble. Now, carefully solder the two pieces together. I expected to need a torch, but a 56-watt soldering iron worked nicely. After a smooth bead is applied around the outside, apply a little solder to the inside of the adapter. This will result in a strong, waterproof joint. Now allow this assembly to cool. After it cools, remove any flux residue to prevent corrosion.

The next step is preparation of the cable itself. Use a tubing cutter and a hacksaw to square off the end.

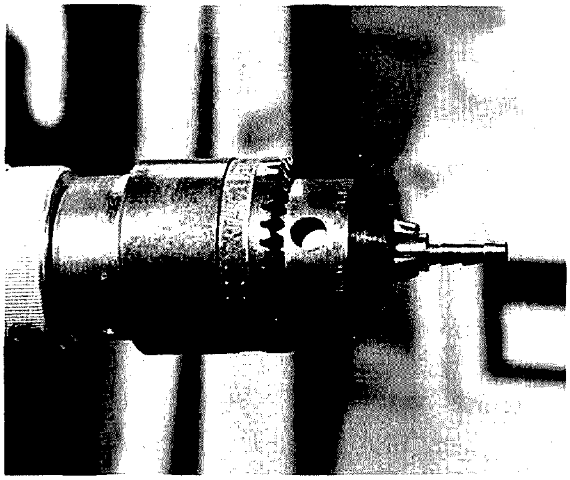

fig. 3. The Barb fitting after one end has been machined. The ribbed end is about to be machined.

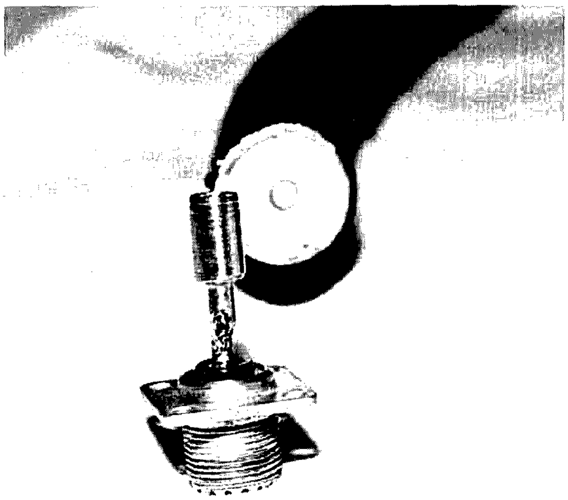

fig. 4. The inner assembly has been completed and prepared for insertion into the outer adapter.

Use the tubing cutter to remove 1 inch (25.4 mm) of the outer insulation. File down the aluminum shield to an outside diameter of 15/16 inch (23.81 mm). Cut the entire cable so that 5/8 inch (15.88 mm) of the cable extends beyond the outer insulation. Carefully square off the center conductor with a fine file. Use the tubing cutter to remove 1/8 inch (3.18 mm) of the shield. Using a sharp knife, cut away the insulation. Do this carefully to avoid nicking the center conductor. This careful order of steps prevents any aluminum filings from contaminating the dielectric. You will now have 1/2 inch (12.7 mm) of the shield extending beyond the outer jacket, and a center conductor extending 1/8 inch (3.18 mm) beyond that. **Fig. 5** shows the completed connector and the prepared cable, ready for assembly.

To place the connector on the cable, carefully start threading the fitting onto the cable. Make sure the fitting goes on square. (A pipe die of the proper size will make this easier, if you can obtain one.) Once the threads are started, you can use a pipe wrench to hold the cable, and an open-end wrench or channel-lock pliers to turn the connector. Do this carefully to make sure you don't kink or bend the cable. Continue screwing the connector on until you feel an increase in resistance. This will indicate that the center fitting has mated. Now carefully remove the connector. Check for stray aluminum filings and any other problems. **Fig. 6** shows the completed connector placed on the cable.

Since there are two dissimilar metals in close contact (aluminum and copper), some steps must be taken to prevent corrosion. Liberally coat the cable shield and the inside threads of the connector with Penetrox or some similar anti-corrosion compound. Now reassemble the connector to the cable. (The Penetrox will act like a lubricant.) Use an ohmmeter to verify continuity from one end of the cable to the other and make sure no shorts exist between conductors. If this test is satisfactory, tape over the connector and the line is ready for use.

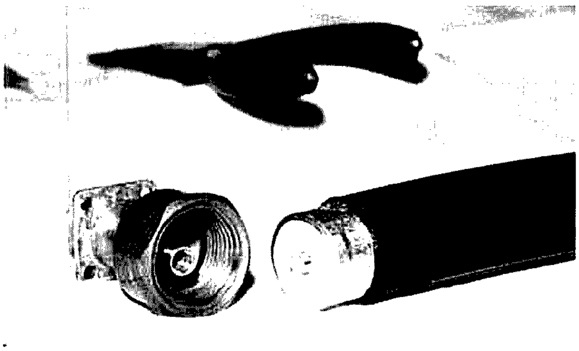

fig. 5. The completed connector and the prepared end of the hardline, ready for assembly.

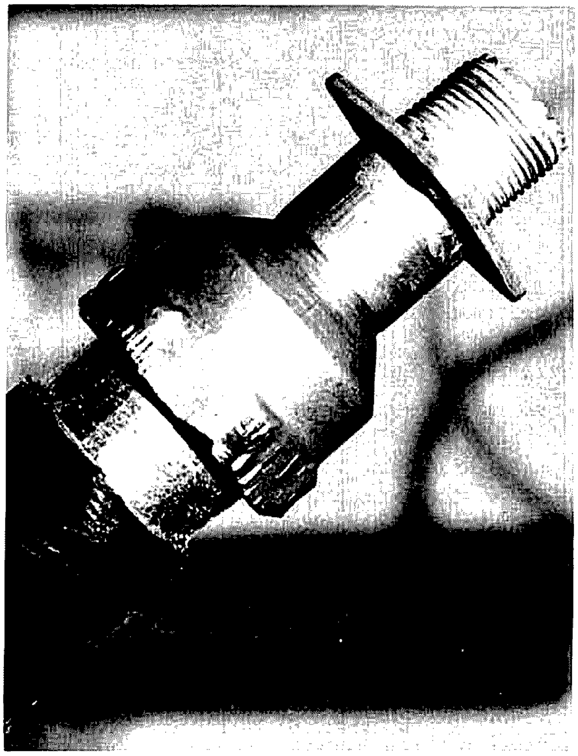

fig. 6. The finished connector installed on the line. It is ready to be protected from the elements and placed in service.

results

The best check of a connector and line assembly is to measure the rf loss through the cable. I tested a 100-foot (30.48-meter) section at 2 meters. The loss measured as 0.8 dB — exactly what the reference tables call for. In other words, the homebrew connectors *did not add any significant loss* to the system.

I have described an economical way to make connectors for 1-inch (25.4-mm) CATV hardline. They are not hard to make, and the materials and procedure can be varied to suit local supplies. Being able to use this high-quality, low-cost cable will make a significant improvement in any station.

acknowledgments

Special thanks go to Mel, W4MJJ, and George, WD4ORM, for their assistance in this project.

ham radio

technical forum

Welcome to the *ham radio* Technical Forum. The purpose of this feature is to help you, the reader, find answers to your questions, and to give you a chance to answer the questions of your fellow Radio Amateurs. Do you have a question? Send it in!

Each month, our editors will select the best answer received to a question posed in the Technical Forum. We will send the writer a book from our Bookstore as a way of saying thanks.

helical antenna matching

In the March, 1983, Technical Forum, a question was raised as to a method of matching a 140-ohm helical antenna to a lower impedance line. A similar problem was covered in the *IEEE Transactions on Antennas and Propagation*, Vol. AP-25, No. 6, November, 1977, Page 913. The antenna design note covers the method of lowering the impedance of the helical to 50 ohms. The method described would appear to be usable at 70 ohms or any other impedance through 140 ohms. — John Belliveau.

Ed note: Most technical libraries probably have files on *Transactions on Antennas and Propagation*.

ham radio thanks Alfred Resnick, K9PXR/9, for his similar solution to the matching problem. In addition he illustrates how series section transformers can be used to transform 70 ohms to 50 ohms. Articles have appeared on that subject in many magazines. Here are some of the sources:

1. Frank Regier, "The Series-Section Transformer," *Electronic Engineering*, August, 1973, page 33.
2. Frank Regier, "Impedance Matching with a Series Transmission Line Section," *Proceedings of the IEEE*, July, 1971, page 1133.
3. B. Bramham, "A Convenient Transformer for Matching Coaxial Lines," *Electronic Engineering*, January, 1961, page 42.

mysterious spur on 160

A local (0.67-mile-distant) 1500-kHz, 50-kW, a-m broadcast station recently installed a new transmitter that uses asymmetrical modulation (95 percent down, 125 percent up). In addition to increasing an already strong rf field, the new transmitter introduced a low-level, broad spurious signal in the 160-meter band that is present on three different receivers. On a sideband receiver the signal is a broad splatter in sync with the station program. On an a-m receiver the signal is intelligible audio.

The transmitter has been cleared by the FCC in response to telephone-equipment-interference complaints. I've estimated the 160-meter "spur" at my location to be about 100 dB down from the 1500-kHz signal. The station engineer was unable to detect it three miles from the transmitting antenna. The spur is difficult to detect closer to the station, but at my location, with a quarter-wave inverted-L, an antenna tuner, and two 1500-kHz traps in the input of the Omni-D receiver, it is an interfering signal of approximately 80 microvolts.

For the first few months the spur seemed to drift randomly in the lower 25 kHz of the 160-meter band over periods of hours and days. When really cold weather occurred in January, I realized that the frequency drift was related to outdoor temperature. Since then I have been correlating the frequency of the spur and the outdoor temperature. A plot of these readings shows that as the temperature rises during the day the spur frequency decreases. The frequency in the early morning is related inversely to the low temperature reached during the night.

Has anyone experienced a similar situation, or does anyone know what is causing this effect? — Jack Geist, N3BEK.

DX FORECASTER

Garth Stonehocker, KØRYW

last-minute forecast

The higher frequency bands (10-30 meters) are favored for the best DX the first half of the month. The solar flux is expected to be highest at that time and lowest about the 20th. Look to the lower frequency bands (40-160 meters) for the best DX the last half of the month. Short-duration disturbed conditions (geomagnetic-ionospheric storms) are expected around the 4th, 12th, and 30th, with a longer-duration event just prior to the 20th. Hearing and working DX will be more difficult during the disturbances, but DX from unusual locations may appear in the form of weak fading signals.

The lunar perigee and full moon, of interest to moonbounce DXers, occurs on the 16th and 26th of this month. An Aquarid meteor shower of interest to meteor-scatter and meteor-burst DXers peaks between May 4th and 6th with rates of 10 and 25 per hour for the Northern and Southern Hemispheres, respectively.

sporadic-E propagation

One of the major paths for excellent DX signals in the summer is short skip, or multiple short skips, on the higher frequency bands. In order to best use sporadic-E (E_S) short-skip propagation, which intensifies toward the end of May and ends in mid-September, a short review is in order: E_S is a thin layer of intense ionization about 60 miles (100 km) above the earth. It gives rise to strong, mirror-like signal reflections over the short-skip distances of 600 to 1200 miles (1000 to 2000 km). Signals remain strong for from a half-hour up to a couple of hours, on the average; they're generally stronger than long-skip. Station location also determines how strongly the present sunspot number (SSN-75) affects *sporadic*-E propagation, with mid-latitudes the least affected and equatorial and polar paths the most. The highest *frequency* propagated by E_S occurs at local noon, since it follows the sun across the sky. However, the highest probability of *occurrence* is near sunrise and again around sunset. These two characteristics of E_S affect short-skip openings differently. Openings on the higher-frequency bands occur near local noontime; the lower bands tend to have openings near sunrise and sunset.

Let's look at the best locations for these E_S openings: Since E_S is related to the summer sun, the effect is in the Northern Hemisphere from June through September and in the Southern Hemisphere during their summer, December through March. The best E_S is on either side of the geomagnetic equator; it's especially good where the geomagnetic equator is furthest from the geographic equator. These special areas are Southeast Asia in the Northern Hemisphere and South America in the Southern Hemisphere. The first is the better of the two.

To look for E_S openings on the higher-frequency bands, monitor beacons on 6 and 10 meters and CB channel 19. Also check TV channels 2 through 5 for 6- and 2-meter openings. The lower bands don't need beacon monitoring since E_S openings (sunrise and sunset) are available most nights.

band-by-band summary

Six meters will provide occasional openings to South Africa and South America around local noontime by short-skip E_S. Monitor TV, an unused channel (2 through 5) for clues.

Ten and fifteen meters will have a few short-skip E_S openings, and long skip during high solar flux to most areas of the world during daylight. Some trans-equatorial openings associated with disturbed ionospheric conditions may occur in the evening hours.

Twenty and thirty meters will have DX from most areas of the world during daylight and into evening almost every day, either long skip to 2500 miles (4000 km) or short-skip E_S to 1250 miles (2000 km) per hop. The length of daylight is now approaching maximum, providing many hours of good DXing.

Thirty, forty, eighty, and one-sixty meters are the night DXer's bands. On many nights 30 and 40 meters will be the only usable bands because of thunderstorm QRN, but signal strengths via short-skip E_S may overcome the static when E_S is available. Although E_S is scarce in May, it should be plentiful next month.

ham radio

MAY

WESTERN USA

GMT	PDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	5:00	20	20	20	10	15	10	10	15
0100	6:00	20*	20	20	10	15	10	10	15
0200	7:00	15	20	20	10	15	10	10	15
0300	8:00	20	20	30	15	20*	10	10	15
0400	9:00	20	20	30	15	20	10	10	15
0500	10:00	20	20	20	15	20	10	10	15
0600	11:00	20	40	20	15	20	10	15	15
0700	12:00	—	40	20	20	20	15	15	20
0800	1:00	—	—	20	20	20	15	15	20
0900	2:00	—	—	—	20	20	15	15	20
1000	3:00	—	—	—	—	20	15	15	20
1100	4:00	—	—	—	—	30	15	15	20
1200	5:00	—	—	—	—	—	15	20	—
1300	6:00	—	—	—	20	—	—	20	—
1400	7:00	—	—	20	15	—	—	—	—
1500	8:00	—	20	15	15	—	—	—	—
1600	9:00	20	20	15	15*	—	—	—	—
1700	10:00	20	20	15	15*	—	—	20	—
1800	11:00	20*	20*	15	10	—	15	20	—
1900	12:00	—	20*	15	10	20	15	15	—
2000	1:00	20	20*	15	10	15	15	15	15
2100	2:00	20	20*	15	10	15	10	15*	15
2200	3:00	20	20*	15	10	15	10	10	15
2300	4:00	20	20	15	10	15	10	10	15

MID USA

GMT	MDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)	CDT
0000	6:00	15	20	15	10	15	10	10	15	7:00
0100	7:00	15	20	15	10	15	10	10	15	8:00
0200	8:00	15	20	20	10	20*	10	10	15	9:00
0300	9:00	15	20	30	15	20*	10	15	15	10:00
0400	10:00	20	20	30	15	20*	10	15	15	11:00
0500	11:00	—	20	30	15	20	15	15	20*	12:00
0600	12:00	—	20	20	20*	20	15	20*	20	1:00
0700	1:00	—	20	20	20	20	15	20	20	2:00
0800	2:00	—	40	20	20	20	20	20	20	3:00
0900	3:00	—	—	30	20	20	20	20	—	4:00
1000	4:00	—	—	—	—	20	20	20	—	5:00
1100	5:00	—	—	—	—	—	20	—	—	6:00
1200	6:00	—	—	—	20	—	—	—	—	7:00
1300	7:00	—	—	15	20*	—	—	—	—	8:00
1400	8:00	20	20	15	15	—	—	—	—	9:00
1500	9:00	20	20	15	15	—	—	—	—	10:00
1600	10:00	20	15	15	10	—	—	20	—	11:00
1700	11:00	20	15	15	10	—	—	20	—	12:00
1800	12:00	20	15	15	10	—	15	20	—	1:00
1900	1:00	20	15	15	10	20*	15	20	—	2:00
2000	2:00	—	15	15	10	15	15	15	20	3:00
2100	3:00	—	15	15	10	15	15	15	20	4:00
2200	4:00	—	15	20	10	15*	15*	15	20	5:00
2300	5:00	—	15	20	10	15	10	15*	20*	6:00

EASTERN USA

GMT	EDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (Caribbean S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	8:00	15	20	15	10	30	10	10	15
0100	9:00	15	20	15	10	30	15	15*	15
0200	10:00	15	20	20	15	30	15	15	15
0300	11:00	20*	20	30	15	30	15	15	15
0400	12:00	20	20	30	15	20	15	15	20*
0500	1:00	20	20	30	15	20	15	15	20
0600	2:00	—	20	20	15	20	15	20	20
0700	3:00	—	20	20	20	20	20	20	20
0800	4:00	—	20	20	20	20	20	20	20
0900	5:00	—	20	20	20	20	20	20	—
1000	6:00	—	—	—	20	—	20	—	—
1100	7:00	—	—	—	—	—	—	—	—
1200	8:00	—	—	—	—	—	—	—	—
1300	9:00	—	15	15	—	—	—	—	—
1400	10:00	20	15	15	15	—	—	—	—
1500	11:00	20	15	15	10	—	20	—	—
1600	12:00	20	15	10	10	—	20	20	—
1700	1:00	20	15	10	10	—	20	20	—
1800	2:00	—	15	10	10	15	20	20	—
1900	3:00	—	15	15	10	15	20	20	—
2000	4:00	—	15	15	10	15	15	20*	20
2100	5:00	—	15	15	10	15	15	15	20*
2200	6:00	20	20	15	10	15	10	15	15
2300	7:00	20	20	20	10	20	10	15*	15

*Look at next **higher** band for possible openings.

Log-Yagis simplified

A 12-foot (boom) antenna achieves 11-dB gain on 10 meters

Several articles on the design of log-periodic dipole and Log-Yagi antennas have made the Amateur fraternity quite conscious of their excellence for long-haul DXing. Their virtues are high gain, exceptional bandwidth, and a large capture area. In order to understand the mathematical concepts, rather than just copying a design, a series of simple functions have been derived that permits any interested Amateur to design his own Log-Yagi.

reflector considerations

Relatively close spacing is employed in these Log-Yagis. Purists may be dismayed by this approach, since approximately 0.5 dB would be lost in a *Yagi* of similar size. In the case of Log-Yagis, however, if such a lost exists it is dwarfed in importance by achievement of front to back ratios of up to 30 to 45 dB. Experimenters who have tried both the wide and close-spaced reflectors report that the close-spaced reflector shows no apparent loss in gain, but that the front-to-back is terrific. Interlacing Log-Yagis does show the loss of about 5 dB F/B when compared with monobanders.

Since I could find no published curves or data for using close-spaced reflectors, I decided to provide my own data at three spacings under 0.15 wavelength. The spacings were chosen to provide easily measured intervals of inches and fractions and result in 0.0765, 0.0854, and 0.1 wavelength. Efficient reflectors are made progressively longer as they are moved closer to the driven element or cell. Simple formulas can then be used to calculate reflector lengths based on the indicated spacing. Finally, the frequencies used for computation are based on the lower band-edge where wavelength is determined by 11808 ÷ f MHz, with the result in inches.

Reflector spacing versus required reflector length is as follows:

spacing	reflector length
0.0765λ	6190 ÷ f MHz
0.0854λ	6115.2 ÷ f MHz
0.10λ	6050 ÷ f MHz

director considerations

In addition to the reflector design needed to produce the best F/B ratio, the best broadband characteristics with constant gain were also considered. Because of perturbations within the log cell, it has been found that with spacings less than 0.12 wavelength the gain is not constant over the entire band. Spacings between 0.125 and 0.150 wavelength exhibit a relatively flat response if the director is adjusted to 95 percent of the longest cell element. The use of spacings of less than 0.125 require pruning or adjusting

By Leo D. Johnson, W3EB, Route 1, Box 448, Hollywood, Maryland 20636

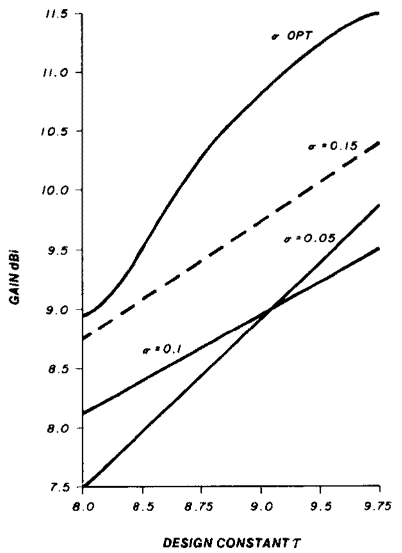

fig. 1. Gain (dBi) vs τ (from K4EWG curves).[1,2,3]

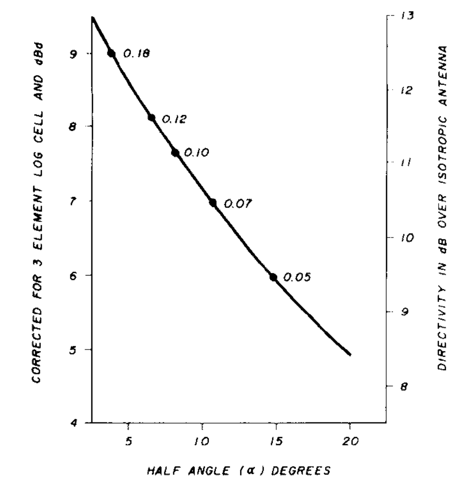

fig. 2. Extension and modification of Isbell's fig. 14.[4] [Curve for $\tau = 0.95$ only with approximate sigmas (σ).]

the director for best results in the portion of the band of interest.

average Yagi gain

Tests conducted using two Yagi parasitic elements with log-cell radiators show 4.3 to 4.6 dB gain over the cell alone. Reference in the text to average Yagi gain is based on a 4.5 dB average.

Second directors provide between 1 and 1.5 dB additional gain when spaced 0.15 to 0.2 wavelength from the first director. A third director seldom adds more than 0.5 dB gain.

the cell function

There are as many combinations of Log-Yagi configurations as imagination will allow. As this article is not a treatise on the construction of a single design, working examples are used to lead the builder through the simple design steps.

In the formulas presented, f is the frequency in MHz at the lower band edge, τ is the design constant between 0.85 and 0.97, and σ is the spacing constant between 0.05 and 0.19 used to determine cell length and gain. Half angle ($\propto$) is the angle formed between the boom and the taper formed by the element.

It should be noted that a τ near 0.95 produces higher gain, with virtually any σ, than is possible using the lower figures near 0.85, and is generally what I use. Bandwidth of the cells, even with high σ, are sufficient through 28 MHz to ensure coverage of the entire band.

Two curves are shown in **fig. 1** and **fig. 2** which enable the designer to reasonably determine cell gain. One represents the τ versus σ from K4EWG's work[1,2,3] and the other is from Isbell's[4] work using τ versus half angles. The Isbell curve has been modified by extending the curves to include half angles near 3 degrees.

Both curves are based on pure log-periodic cell design and their accuracy is not questioned. For Log-Yagi work, Isbell's curves appear to correlate closely if a correction factor of -1.3 dB is applied.

Subtraction of 2.2 dB results in dBd — or gain over a dipole. For this reason, the left-hand figures on the modified Isbell curve have been corrected by 3.5 dB and shown as dBd.

Either curve shows that cell gains over a dipole, when added to the average Yagi gain, *provide a very efficient antenna on a relatively short boom.*

designing the antenna

Having waded through the basics that are pertinent to Log-Yagi design, you can proceed with the development of the antenna shown in **fig. 3** using simple formulas.

For the cell half-lengths in inches:

$$\ell 1 = 2820 \div f$$
$$\ell 2 = 1 \times \tau$$
$$\ell 3 = 2 \times \tau$$

Spacing between the elements is calculated by first multiplying the selected σ by four and again multiplying that quantity by the length of $\ell 1$. Stated as a formula: $\ell 1(4\sigma) = \ell 1 - \ell 2$ spacing. To calculate the

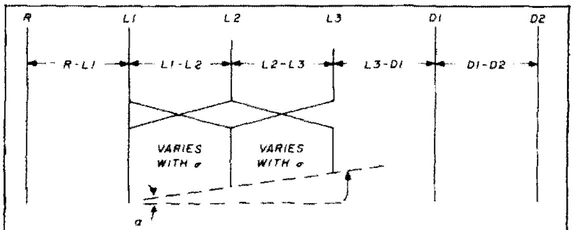

fig. 3. Log-Yagi consisting of cell (ℓ1, ℓ2, and ℓ3) and up to three parasitic elements.

ℓ2 − ℓ3 spacing multiply the ℓ1 − ℓ2 spacing by τ.

This completes the cell design and a total Log-Yagi can be designed from the data presented so far.

For example, a 28-MHz antanna with a τ of 0.95 and using a σ of 0.07 results in the following cell dimensions.

$$
\begin{aligned}
\ell 1 &= 2820 \div 28 = 100.71 \\
\ell 2 &= 100.71 \times 0.95 = 95.6786 \\
\ell 3 &= 95.6786 \times 0.95 = 90.895 \\
\ell 1 - \ell 2 &= (4 \times 0.07) \times 100.71 \\
&= 0.28 \times 100.71 \\
&= 28.1988 = (28.2) \\
\ell 2 - \ell 3 &= 28.2 \times 0.95 \\
&= 26.79 = (26.8) \\
\textit{cell length} &= 55 \textit{ inches}
\end{aligned}
$$

Continuing the design for the parasitic elements using 0.0765-wavelength spacing for the reflector and 0.15-wavelength spacing for the director we find:

$$
\begin{aligned}
R &= 6190 \div 28 \\
&= 221.07 \\
R - \ell 1 &= (11808 \div 28) \times 0.0765 \\
&= 421.7 \times 0.0765 \\
&= 32.26 = (32.25) \\
d &= (2 \times 100.71) \times 0.95 \\
&= 201.42 \times 0.95 \\
&= 191.349 = (191.35) \\
\ell 3 - d1 &= 421.7 \times 0.15 \\
&= 63.25
\end{aligned}
$$

The parasitic elements require 95.5 inches plus 2 inches each for mounting; when added to the cell length, this figure indicates that a boom of 154.5 inches, or 12.875 feet, is required. If the antenna was to have been designed for exactly a 12-foot boom, then this example must be changed by reworking the cell length or changing the director spacing. In the example given, reducing the director spacing to 0.125 wavelength results in a new spacing of 52.75 and the antenna fits a 12-foot long boom nicely.

The K4EWG curve indicates a cell gain of 9.2 dBi, or 7.0 dBd. To compute the half angle to check with the modified Isbell curve, we must calculate the cotangent (cot) of the half angle from the τ and σ used in our design as follows:

$$
\begin{aligned}
\cot \alpha &= (4 \times \sigma) \div (1 - \tau) \\
\cot \alpha &= (4 \times 0.07) \div (1 - 0.95) \\
&= 0.28 \div 0.05 \\
&= 5.6
\end{aligned}
$$

Cot 5.6 (5.614) resolves to a half angle (α) of 10.1 degrees.

The gain on the modified Isbell curve indicates 8.8 dBi, or 6.6 dBd, for the cell alone. Cell gain of 6.6 plus 4.5 average Yagi gain renders a figure of **11.1 dBd total** gain for the Log-Yagi, or about 0.6 dB less than indicated by the other curve.

The two methods produce little difference in cell gain figures in the region between sigmas of 0.05 and 0.12, but even the lowest of gain figures equates to a power ratio of 12.6, which makes 100 watts as effective as 1.25 kW on a dipole.

wide-spaced cells

The previous design produced a high-gain antenna on a short boom. Surely some designers will be considering whether versions with longer booms and more directors are practical, particularly for those who have the space to erect them.

If all the constants remain the same except σ, which is increased, only the spacing between cell elements will change. The spacing for ℓ1 − ℓ2 becomes 68.5 inches and ℓ2 − ℓ3 is 65.063 inches for a cell length of 133.5 inches using a σ of 0.17.

Using this cell length with 0.15-wavelength director spacing and 0.0765-wavelength reflector spacing, the boom required would be a little over 19 feet long. If, however, the reflector spacing were changed to 0.0854 wavelength, the mechanical balance would be improved and the configuration would fit nicely on a 20-foot boom.

Using the previous formulas, the $\cot \alpha$ is 13.6 and the half angle is 4.2 degrees. The modified Isbell curve shows a cell gain of 8.95 dBd and a total Log-Yagi gain of **13.45 dBd**. The 100 watts now looks like 2 kW on a dipole.

While straining for every dB possible, adding a second or third director could give a final figure of over 15 dBd.

tolerances

Two items left untouched by most other articles on this subject are the need for careful workmanship and the use of relatively finite measurement if the best results are to be attained. Inattention to detail or poor workmanship can cost you gain.

Tolerances should be held to 1/16 inch for element

lengths and spacings up to 1/8 inch as high as 28 MHz. For metric measurement, 1 mm is an excellent tolerance figure (for both length and spacing).

By fastening the phase lines exactly 0.5 inch from the attachment end of the radiator, and maintaining equal lengths of each wire or strap in the phasing pairs, the builder is ensured of good electrical balance and his results will be repeatable time after time. The dimensions developed from the design effort are based on center-to-center spacing of all elements.

fine tuning the design

In many combinations of the three basic factors of design, it appears that some fractions make the measurement practically impossible. Other cases are noted where attaining the tolerance figures for construction is impossible.

Changing one or more of the factors even slightly can often resolve the problems. In the following example of a 14-MHz design, the original figures and finalized computations are explained:

original computation	final computation
$f = 14\ MHz\ \ \tau = 0.95$	$f = 14.0037214\ \ \tau = 0.950341403$
$\sigma = 0.1791$	$\sigma = 0.1789265$
$\ell1 = 201.42857$	$\ell1 = 201.375$
$\ell2 = 191.357$	$\ell2 = 191.375$
$\ell3 = 181.7893$	$\ell3 = 181.875\ (181.8716)$
$\ell1 - \ell2 = 144.303$	$\ell1 - \ell2 = 144.125$
$\ell2 - \ell3 = 137.088$	$\ell2 - \ell3 = 137.0\ (136.968)$

First, the dimensions of $\ell1$, $\ell2$, and $\ell3$ were difficult to measure. This was resolved by dividing 2820 by 201.375 for the new frequency. Although $\ell2$ and $\ell3$ could be considered within tolerance, it was desirable to see how τ would be influenced.

The figure of 191.3786 for $\ell2$ after the frequency was changed was close to 191.375, so a new τ was developed by dividing 191.375 by 201.375 for $\tau = 0.950341403$, which helped make $\ell3$ a more easily resolved figure.

Although the cell spacings were resolvable, I felt that reducing the sigma slightly would permit the use of integral inches for $\ell2 - \ell3$, and that the small change would not affect gain. By cut and try, I improved the dimensions and arrived at the new figure.

The results are dimensions well within the established tolerances. It is much more simple to redo the arithmetic than to try to measure uncommon fractions!

construction

I've tried various methods for mounting cell elements. Generally, the insulating material used in cell construction dictates the mounting method. When using polystyrene, Lucite, Plexiglass, or PVC tubing as insulators, strap them with stainless steel hose clamps. (If you use U-bolts, a cushioning material must be added.) With these insulators, I used 1-1/4 × 1-1/4 aluminum angle mounted to 4 × 4 plates for fastening to the boom (with muffler clamps). Most of the materials mentioned succumb to weathering of some sort in two to three years. PVC shows breakdown of insulation and the others get brittle and crack.

The best material is polycarbonate. Though this material is expensive, it has a tensile strength of 6000 psi, a breakdown characteristic of 360 volts per mil (0.001 inch), it retains its impact strength to −40 degrees F, and it has a temperature distortion point of over 260 degrees F. Polycarbonate with 1/8-inch wall can support a full-sized 14-MHz element, with two U-bolts spaced 6 inches apart, when the element is enclosed in a tube only 7 inches long with a gap between elements ends of 0.5 inch. There will be no noticeable sag at the element center.

guying

Single guy wires are satisfactory for small booms and on larger-diameter long booms with thick walls. The extra support provided by umbrella-type guying is recommended in most other cases. When the installation is close to salt water, or in areas where oxidation levels are high, stainless steel guys and turnbuckles are highly recommended. The 3/32-inch sailboat-shroud cable is adequate for most cases. For very heavy arrays, such as interlaces, 1/8-inch material is recommended. Dacron is the only rope material recommended for guys. This should be of the woven type, in diameters of 1/4 or 5/16 inch. Rope guys increase wind resistance considerably.

matching

Impedances of almost all configurations are between 35 and 48 ohms. Whether strap, rods, tubes, or wire is used for the phasing lines, their influence is small so far as matching capabilities are concerned.

K4EWG devised a matching stub for his design which is easily found by using 256 ÷ f. It is installed between $\ell3$ and a 1:1 balun. Closing up the stub spacing or adjusting 1/8 inch at a time provides the best match.

On many occasions it is difficult to make such changes easily. A preferred method is to feed the antenna through a balun and slightly shorter stub, using a transformation in the feedline. This approach uses either an odd number of quarter wavelengths of 50-ohm feedline (corrected for velocity factor) or a single 50-ohm quarter-wave section between 70-ohm

feedline and the balun end. In every case, it has been possible to reduce the VSWR to 1.3:1 or less across the band.

Though the standard computation for a quarter-wavelength section is 246 ÷ f multiplied by the line's velocity factor, my figure of 240 ÷ f low end, with correction for velocity factors, appears to provide the best broadband characteristic for this transformation.

about gain figures

I will be among the first to agree that antenna models do not guarantee the gain figures attributed to the designs of these Log-Yagis. Usually as much as 2.5 dB variation is noted.

Recent antenna testing by the NRL (Naval Research Lab) and others have verified that modeled antennas are but guidelines for the development of full-scale antennas, where certain performances are required over particular paths. Recent testing has been performed with full-scale antennas to determine "apparent gain" over such paths.

The use here of the idea of apparent gain is similar to its use in the development of "gain type" antennas for mobile use. For example, a 5/8-wavelength antenna by itself cannot produce a 3-dB improvement over an antenna 1/4-wavelength long. Apparent gain is accomplished by concentration of energy in a favorable direction or takeoff angle.

Apparent gain follows the design gain quite closely in Log-Yagi arrays. On long-haul paths of over 3000 miles, a comparison with a reference dipole yields results that are quite close to those derived by computations using the curves. The large capture area and non-symmetrical vertical pattern are no doubt contributors to its ability on such paths.

credits

Thanks goes to Peter Rhodes, K4EWG, for planting the original seed and for taking the time to discuss and verify the aspects of this new design; to WA3ELE for making the first long-boom wide-spaced array; and a special thanks to the model shop-workers who manufactured the antenna hardware.

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ham radio

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the grounded monopole
with elevated feed

Low-takeoff-angle vertical for 10 through 80 meters

A popular multiband antenna in the 1930s was the off-center-fed or Windom. It consisted of a half-wave horizontal dipole, at its fundamental frequency, fed off-center by a single wire feeder, at a distance of about 0.36 times its length measured from one end. A later version of the off-center-fed antenna (1940s) used 300-ohm twin lead instead of a single wire feeder, fed at a point one-third of its length, measured from one end. This antenna operates satisfactorily on the fundamental frequency and on harmonics, permitting operation on the 80-, 40-, 20-, 15-, and 10-meter bands.

If this off-center-fed antenna is turned up on end and grounded at the end closest to the feed point, we now have a vertically-polarized antenna with impedance and radiation characteristics that change with frequency in such a way that this antenna can be successfully employed for multiband (multi-frequency) operation. However, it has not been used, to my knowledge, at high frequency for radio communications. The antenna is in effect a grounded vertical monopole with elevated feed. Its main lobe, which is directed toward the horizon, does not break up into a high angle lobe for heights between 3/4 and 1 wavelength.

The transfer of the feedpoint from the base upwards has been used for a different purpose, in the sleeve antenna, originally designed for VHF, but recently adapted for use at high frequency.[1] The half-sloper[2] is also a type of elevated feed antenna. In the present design, the antenna is earthed at its base, and sectioned at a height of one-third its total height. The coaxial feeder cable is brought up along or inside the earthed lower section. Its sheath is connected to the top of the lower section, and the inner conductor is connected through a 4:1 step-up transformer to the insulated upper section. This is shown in **fig. 1A**.

An antenna of this design was described by Hatch et al[3] who analyzed it by approximation, treating the antenna as a lossless transmission line of constant characteristic impedance. Since the standing wave component of the antenna current is much larger than the progressive wave component, corresponding to radiation, for thin monopoles, this treatment is a good first approximation.

On this assumption, the authors computed the current distribution on the radiator for $h = \lambda/4$, $\lambda/2$, $3/4\lambda$ and λ, (**fig. 2**). Note the elevated feed has a pronounced effect on the current distribution on the radiator, an effect which improves the radiation pattern of the antenna for $h > \lambda/2$, since for $h = 3/4\lambda$ and λ the current distribution is essentially in phase, a desirable feature for maximum gain.

radiation patterns

The radiation patterns of the monopole were also computed, and are reproduced in **fig. 3**, for $h = \lambda/2$, $3/4\lambda$ and λ. Patterns for the elevated feed differ little from those for base feed for heights up to $h = \lambda/2$, but there is a substantial improvement in low angle

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radiation for $h = 3/4\lambda$ and λ. In the case $h = \lambda$, the base-fed antenna has only a high angle lobe, whereas with an elevated feed, there is no high angle lobe, and the radiation is dominantly low angle (less than 10 degrees above the horizon). Such an antenna would be a good DX antenna since it will have gain at these frequencies. The patterns are significantly modified by the finite conductivity of the earth, and a radial ground system must be employed to reduce losses due to currents returning to the base of the antenna through the ground. This is no different from any ground plane antenna.

antenna reactance

The reactance to the source was computed, and calculated curves are reproduced in **fig. 4**. The rate of change of reactance with frequency is smaller for the elevated feed antenna, and the SWR (actually X/Z_0, where Z_0 = the characteristic impedance of the antenna if considered to be an open-circuit transmission line) is particularly *small at* $\lambda/4$ and $5\lambda/4$. The SWR at $\lambda/2$ and λ is acceptable if an antenna tuner is used to match the antenna to the transmitter. If the antenna height is such that it is approximately quarter-wave resonant at 80 meters (3.75 MHz), it could be used on 80-, 40-, 20-, 16-meters (18 MHz) and 15 meters ($h = 3\lambda/2$).

antenna modeling

The antenna reactance versus frequency curve shown in **fig. 4** represents the ideal case, since the antenna was analyzed as a lossless transmission line, whereas a practical antenna has resistance (radiation and loss resistance) as well as reactance. The impedance of a modeled antenna over a perfectly conducting ground plane (a 30-meter diameter wire grid ground screen) is shown in **fig. 4**. The antenna was 1.96 feet (60 cm) high, and 0.25 inches (0.63 cm) in diameter. The antenna was fed at a point one-third its height above ground. The top section was fed by connecting it to a wire running inside the lower section, but insulated from it. The lower section of the mast with the feed wire inside behaved like a coaxial feeder, as well as part of the monopole antenna. The impedance (Z, θ and Γ) was measured between the lower end of the coaxial feeder wire and ground. Γ is the voltage reflection coefficient, with reference to 50 ohms:

$$\Gamma = \frac{SWR - 1}{SWR + 1}$$

If at full scale 3.5 MHz corresponds to 100 MHz, the scale factor equals 28.57, and at full scale the monopole is 56.24 feet (17.14 meters) high, and 7.4 inches (18.14 cm) in diameter. For this scale factor, the band edges for the 80, 40, 30, 20, 15, and 10 meter bands are marked. Except at 20 and 40 meters the $SWR < 4{:}1$.

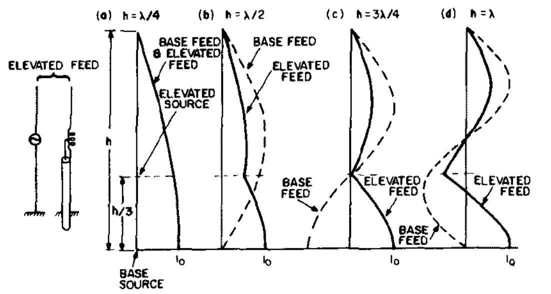

fig. 2. Current distribution for base-fed and elevated-feed monopoles.[3]

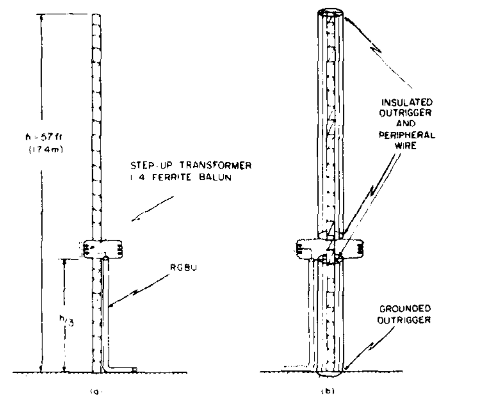

fig. 1. (A) Elevated feed grounded monopole antenna, (B) Use of a cage of wires instead of physically sectioning the antenna.

sectioning the monopole

A tower is physically sectioned by proper placement of insulating sections. This is not very practical, especially if a grounded tower is available. Broadcasters have used grounded towers for particular applications that require the tower to be sectioned, and they have devised a method to effectively achieve this without *physically* doing so. The method is sketched *in* ***fig. 1B***.

The tower is screened using insulated outriggers which support a surrounding cage of vertical wires. Six or eight wires are required, although four wires, as sketched, might be satisfactory. The wires are joined together by a peripheral wire at the top of the

tower, at the bottom of the top section, at the top of the bottom section, and at the base of the tower. The sketch shows a physical separation at the place where the tower is sectioned. In practical applications, a series strain insulator would be inserted in the vertical dropwire at that point. This arrangement effectively screens the grounded tower, sections it, and since the electrical diameter is increased, the intrinsic bandwidth of the radiator will be greater.

performance

A temporary test antenna was constructed using a 37-foot free-standing whip mounted on an 18-foot lattice tower. This antenna was erected at the author's QTH (**fig. 5**). The SWR was measured at a number of frequencies in 3-30 MHz band. These results are plotted in **fig. 6**, where the abscissa is h/λ rather than frequency.

Since the antenna is not resonant and matched at any frequency in this band, the SWR depends upon the length of the feeder transmission line, and its characteristic impedance. The SWR for lengths 30 feet and 100 feet of RG8-U (50-ohm coax) was measured, and measurements were made with 72-ohm coax. Rice and Winacott,[4] following the Marconi work, employed a 7.5 μH coil across the 4:1 step-up transformer, (**fig. 1**), which was supposed to improve the SWR at the higher frequencies. The author found that this coil increased the SWR at these frequencies, and so this inductor was not used. While there were differences in the SWR at particular frequencies for the different lengths and impedances of the feeder cable, an optimum length or impedance was not found. The results in **fig. 6** were for a 100-foot length of RG8-U. The SWR for the various present and proposed Amateur bands are in **table 1**.

The SWR was highest at 10.1 MHz, where $h/\lambda = 0.57$, it was 5.5. This is, however, of no consequence, provided the antenna can be matched employing an antenna tuning unit. Since the normal loss for $SWR = 1$ for RG8-U cable at 10 MHz is 0.45 dB,

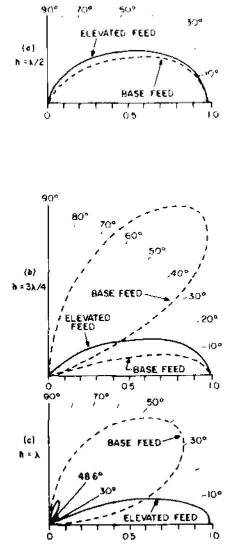

fig. 3. Vertical radiation patterns for base-fed and elevated-feed monopoles.[3]

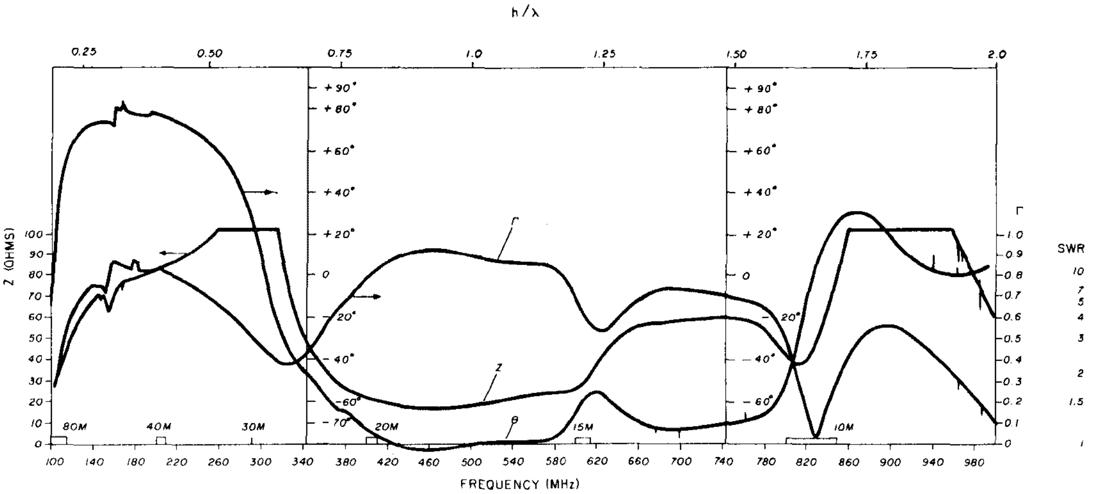

fig. 4. Curves that show the feed impedance of the grounded monopole with elevated feed versus operating frequency for the scaled model. Corresponding full-scale frequencies are noted.

table 1. Initial antenna SWR measurements are indicated together with the corresponding Drake MN-4 antenna tuner dial settings and bandwidth positions. The last column lists impedances inferred from the tuner R/X dial settings.

frequency MHz	initial SWR	antenna tuner dial R/X	antenna tuner bandswitch	antenna impedance magnitude (ohms) and phase (degrees)
3.70	3.3	9/1.1	80	26 / 30
7.10	4.5	2.1/3.6	40	98 / +46
10.10	5.5	5.6/7.8	40	100 / +59
14.15	3.3	4.6/4.1	20	17 / 34
18.10	2.0	4.3/8.6	20	85 / +34
21.20	2.8	4.6/5.3	15	19 / +16
24.90	2.5	4.1/10	15	83 / +15
28.50	3.1	3.9/7	10	19 / 50

the additional loss due to $SWR = 5.5$ is about 0.6 dB, for a total loss of about 1 dB. This is hardly worth worrying about. A more important consideration is radiation from the transmission line, which should be buried, and the run above ground into the shack should be as short as possible.

Table 1 includes dial setting and bandswitch positions to tune the antenna at the various frequencies for an $SWR = 1:1$, employing a Drake MN-4 antenna tuner. Also shown are the antenna impedances inferred from these dial settings. That is, the transmitter input port was terminated in 50 ohms and with the tuner controls reset to the indicated value, the conjugate impedance was measured at the tuner output port. This measurement gives the correct magnitude of the antenna impedance, but the opposite sign of the phase angle. These settings and impedances would not apply to other installations, since it is hardly likely that this temporary antenna would be copied identically. They are given to indicate that the antenna can be tuned at all frequencies using an antenna tuner. A table such as this facilitates band change, the controls can be preset and require only trimming for minimum SWR.

While I had no doubt that the antenna would perform as predicted, my concern was that losses in the ferrite balun (which was used for the 4:1 step-up transformer) might be high for high SWR.[5] I do not have a Drake B-1000 balun which is supposed to be designed for such applications. For outdoor use, this balun must be mounted in a weatherproof box, with feedthrough insulators.

The first test was to measure the SWR at different power levels. It was measured at 10 watts of forward power and 100 watts of forward power. No difference was detected.

The operational performance of an antenna is difficult to measure quantitatively. The following account describes some communications tests conducted over several days in October, 1980.

Starting with 20 meters, I measured the relative gain with respect to an elevated ground plane (a Hy-Gain 14AVQ trap vertical with 16 radials, four for 40, 20, 15, and 10 meters) on the roof of my garage. A gain of 0 to 1 S units was measured (0 to 5 dB). The measured gain was obviously dependent on the distance of the station being received and the propagating mode (angle of elevation of signal received).

On 40 meters during early evening hours, I worked UK2PCR, and GW4BWK, whom I chatted with for half an hour or so. He was using a full-wave delta loop apex up, lower corner feed (vertical polarization). I was using a Yaesu FT101 (100-watt transceiver).

On 75 meters, during the same two evenings, I worked Y21UJC, EA1UU, FT7DG, and G2PU. I had not previously worked DX from my QTH on 75 meters, since my fixed antenna system is quite inadequate for working DX. If you can't hear DX, you can't work it.

I QSY'd with VE8MA from 20 meters to 15 meters late one evening. I thought the 15 meter band might be dead. My received signal report came up by an S-unit, his remained the same. He was using a tri-band beam.

On 10 meters, my brief experience is that if you can hear the station you can work him.

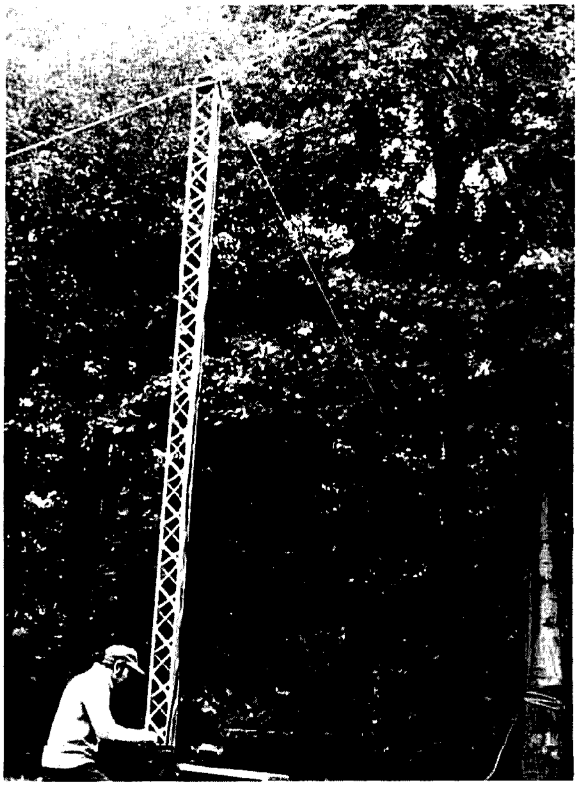

fig. 5. Photograph of a simply-constructed monopole with elevated feed (employing a free-standing Fiberglass whip for the top section).

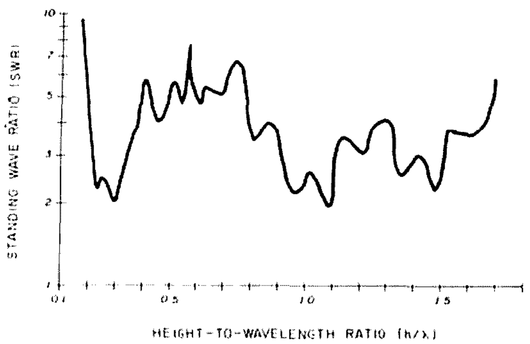

fig. 6. SWR versus height/wavelength ratio for the antenna shown in *fig. 5*, fed with 100 feet of RG8-U.

conclusions

The antenna appears to perform well. It is not a gain antenna, and beam antennas usually, but not always, outperform it. The lack of a directional pattern means that QRM can be high. However, the grounded monopole antenna with elevated feed has a pattern and impedance that changes with frequency in such a manner that the antenna can be used for DX, and it can be used on any frequency in the high-frequency band (3 to 30 MHz).

appendix

Radio Amateurs nowadays are accustomed to employing matched antennas, and some might find it difficult to match their transceiver to a reactive load. As an aid:

1. Calibrate the dial settings for your transceiver using an SWR bridge and a 50-ohm load;

2. when *tuning a reactive antenna* (high SWR) don't tune the transmitter PA for maximum forward power to the mismatched antenna. You will only mistune your transmitter. Set the plate tank and load capacitors to the place where the transmitter delivers maximum power at an $SWR = 1:1$ into the 50-ohm load;

3. with low rf drive (sufficient to measure SWR or reflected power, tune the resistance and reactance dials of the antenna tuner together for low SWR (or reflected power). Only when the antenna is matched, and the SWR seen by the transmitter is 1:1, should you tune the transmitter for maximum forward power.

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4. D.W. Rice and E.L. Winacott, "A Sampling Array for HF Direction-Finding Research," Communications Research Centre, *CRC Report No. 1310*, November, 1977.

ham radio

the ham notebook

Radiokit
Box 411
Greenville, New Hampshire 03048
(Millen capacitors stock No. 19100)

Amp Supply
2071 Midway drive
Twinsburg, Ohio 44087
(Capacitors)

In all instances send SASE for reply.

Ed Marriner, W6XM

another antenna tuner

One type of antenna tuner that has not seen much use is one half of the Johnson Match Box circuit (see **fig. 1A**). The advantage of this circuit

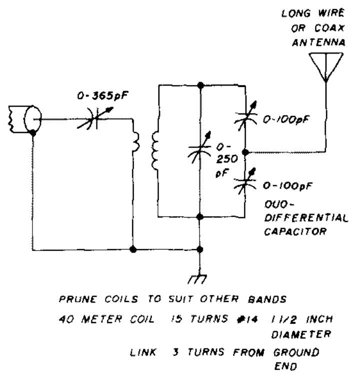

fig. 1A. Antenna tuner schematic.

over some of the more popular transmatch tuners is that it is a resonant circuit that rejects harmonics. It is very easy to adjust for either a long wire or coax-to-coax feed systems. It will not work for open-wire line because it is only half of the original circuit.

The reason it has not been used is probably the difficulty in obtaining a duo-differential capacitor. This capacitor has two stator sections and one combined rotor section. As the rotor increases in one section the other decreases. It forms a capacitance tap for the antenna across the coil in place of sliding the antenna tap around for a match. A duo-differential capacitor can be made by soldering two capacitors together or by mechanically linking two capacitors. The two capacitors are joined by remov-

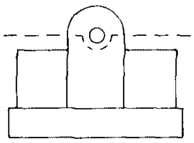

fig. 1B. Millen type 19100 was used here because the rounded top of the support made it easier to form and solder. Other types might also work.

ing the shaft of one, cutting across the middle of the hole, and enlarging the hole to fit on top of the other capacitor (see **fig. 1B**). After fitting, solder as shown in **fig. 1C**.

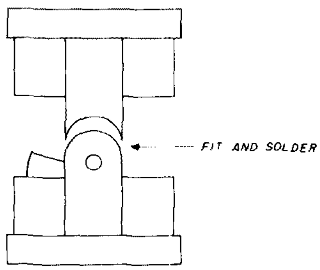

fig. 1C. Join the two capacitors as indicated.

I built my tuner in a box with a five-prong ceramic tube socket mounted on insulators by punching a hole in the top of the box. The coil, manufactured by Bud, was mounted using the five-prong socket. The small BDC type handles several hundred watts and makes a nice small tuner. The tuner controls are adjusted for minimum SWR.

sources

Fair Radio Sales Co.
P.O. Box 1105
Lima, Ohio 45802
(Bud coils, capacitors)

latching relay control

The great virtue of a latching relay is that it draws current only while it is changing states. The latching relay's built-in magnet holds the contacts in their last position until they are changed by a current flow through the relay winding. This makes the latching relay ideal for use with battery-powered or remote equipment.

When the circuit in **fig. 2** is completed through the switch, electrons flow through R1 to the negative terminal of C1, which was in a discharged state (both plates of the same polarity); the capacitor looks momentarily like a conductor. Thus the voltage appearing at the junction of C1/R1 rises to near the supply value, and this surge of electrons flows to Q1's base, causing Q1 to conduct and energize the relay, changing its state.

As C1 becomes fully charged, the electron flow ceases. Therefore Q1 no longer conducts.

The schematic diagram of **fig. 3** uses a similar principle to control a latching relay. CR1's forward resistance guarantees that Q1 will not be

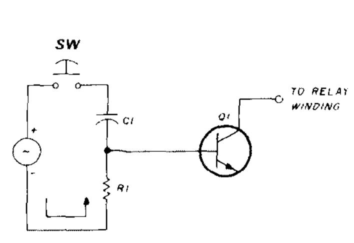

fig. 2. Block diagram of a latching-relay controller.

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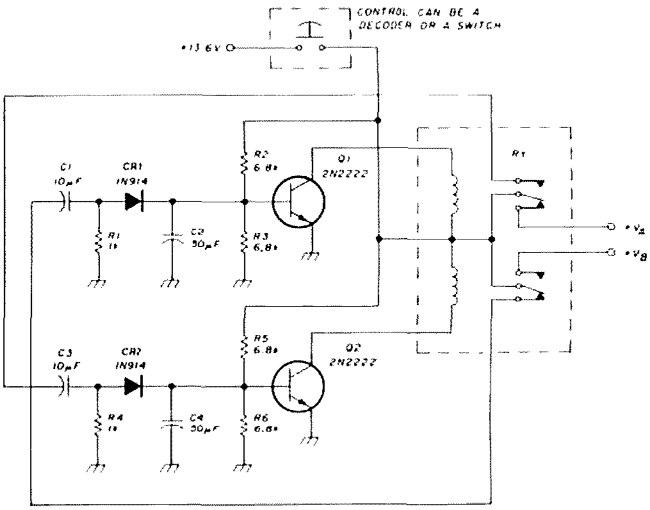
the ham notebook

continued



C1,C3	10-μF
C2,C4	50-μF
R1,R4	1K
R2,R3	
R5,R6	6.8K
Q1,Q2	2N2222
RY	28 volt, DPDT: Potter & Brumfield SL-7156 28 volt, DPDT: General Electric 3SAM1283A2 12 volt, 4PDT: RCA 231435/Gould-Allied Control T3ZX-10

fig. 3. Schematic diagram of the latching-relay controller.

turned on by a voltage of less than 0.6 volt. The contacts automatically complete the circuit to Q2 so that the next closing of the switch will cause the relay to again change state in the manner described.

Latching relays in the 12-volt range are hard to find, but surplus relays rated from 24 volts to 28 volts are common. They will work down to about 11 volts. Surplus relays are usually of the "crystal can" size. The RCA/Gould relay listed has a 12-volt coil and four poles, and it is larger than the others. It is a specialty item that might be obtainable only through RCA suppliers.

An ideal use for this control is an on-off power switch for a battery-operated repeater. A low-current receiver with a decoder IC connected to the control line of this circuit will allow the repeater to be turned on or off on command.

Charles G. Bird, K6HTM
Chico, California

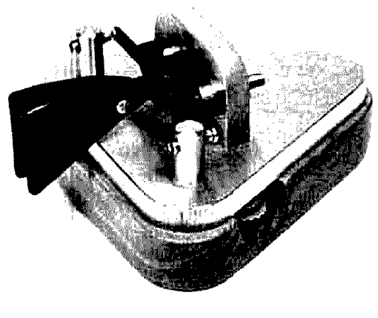
NEW products

Vibroplex electronic keyer

When you think of speed keys, one of the first that comes to mind is the Vibroplex "Bug." Years before the advent of electronic keyers, they were the only alternative to the "cootie key" for sending semi-automatic code. In those days, you could actually tell who was sending well before signing of calls by the sender's fist. Dahs were drawn out longer than they should. Dits came in a rapid-fire "bruuup." Now, however, with keyboards and other forms of electronic keys, everyone sounds somewhat the same.

Vibroplex has introduced a new Iambic key and keyer that will be of great interest. Their new Brass Racer, based on the FYO design, is attractively built and is rock solid. The Vibroplex treatment of this design does not use springs to adjust paddle tension. A clever use of magnets controls the paddle tension.

Another twist is that Vibroplex took the Brass Racer base, hollowed out the center and inserted an electronic keyer that uses the Curtis 8044 chip. This makes for a very nice, compact, self contained keyer (a big plus for field day and other portable operations).



There are no power cords to clutter up the operating desk. Power comes from a self-contained 7.5-volt battery. The EK-1 is limited in that it does not have a memory like so many of the newer electronic keys. But not everyone feels that this is necessary and many will find the EK-1 a nice, simple package.

For more information, contact Vibroplex Company, Attention Bruce Palmer, P.O. Box 7230, Portland, Maine 04112; Reader Service Number 301.

tri-band vertical

Hustler, Incorporated, has announced a three-band vertical antenna for 10, 15, and 20 meter operation. A unique two-in-one trap design allows excellent bandwidth while maintaining an overall height of only twelve feet.

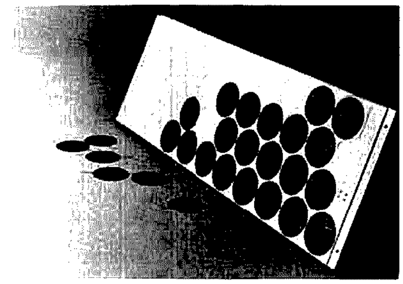
The antenna, 3-BTV, is designed for permanent ground mounting or for portable use on travel trailers, condo balcony railings, or anywhere exhibiting a sufficient groundplane.

The antenna is made of high quality aluminum with stainless steel hardware, supplied with a heavy duty bracket for pipe or bulkhead mounting.

For additional information, contact Hustler, Incorporated, 3275 North B Avenue, Kissimmee, Florida 32741. Reader Service Number 302.

photovoltaic panel kit

Encon announces solar panel kits for the Amateur that enable you to build your own solar electric panel for less than $6.00/watt.



Molded high-strength plastic base has forty 4-inch recesses and thirty-six 4-inch diameter cells. One panel should produce approximately 17-volts, between 1.2 and 2 amps. Cover glass, silicone potting, wire, and solder, not included.

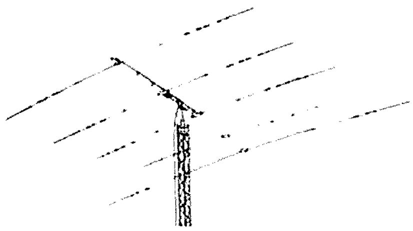
These kits are ideal for demonstrators and schools seminars. Good working panels have been constructed in less than two hours each. Instructions are included; it is recommended that you have basic soldering skills.

For more information, contact Encon Corporation, 27584 Schoolcraft Road, Livonia, Michigan 48150; Reader Service Number 303.

TH5Mk2 tribander

The TH5Mk2 is a five-element broadband tribander for 20, 15, and 10 meters. The TH5Mk2 will load tube-type or solid state auto-tuned rigs from band edge to band edge on 20 and 15 meters. On 10 meters there is a choice of 28.0 to 29.4 or 28.3 to 29.7 MHz, all below 2:1 VSWR. The Hy-Q traps for each band are the most efficient technique for multibanding a Yagi antenna. Factory assembled and pre-tuned traps are mechanically superior, and provide reliable all-weather performance. With four active elements on each band, the average forward gain is an impressive 8.5 dB, and average front-to-back ratio is 20 dB.

The antenna assembles on a 19 foot (5.8 meter) boom. With a maximum element length of 31.5 feet (9.6 meters), the tuning radius is only 18.4 feet (5.6 meters). The assembled antenna weighs 59 pounds (26.8 kg).

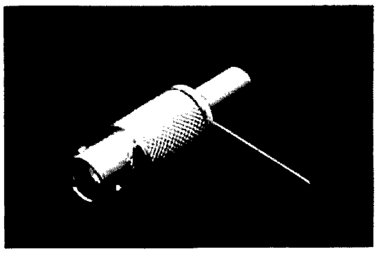


The antenna includes stainless steel hardware, the BN86 balun and a sophisticated matching dual-driven element feed system as also used in the larger TH7DX. The antenna provides dc grounding for lightning protection. The suggested price is $459.95. For more information, contact Hy-Gain, 9600 Aldrich Avenue, South, Minneapolis, Minnesota 55420.

BNC adapters

Centurion International, Inc., has introduced a new line of BNC adapters designed for antenna connection to two-way portable radios that require threaded connectors.

The adapters are available in nine different styles and feature a grounding strap for use with portable gain antennas that require ground potential. The adapters may also be used with mobile antennas, mobile amplifier chargers, and a variety of other applications.



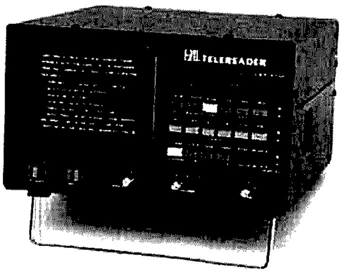
For more information, contact Centurion International, Box 82846, Lincoln, Nebraska 68501. Reader Service Number 304.

receive-only RTTY/CW terminal

HAL Communications Corp. announces the new CWR6750 receive-only RTTY/CW terminal. The CWR6750 is the ideal companion to a shortwave receiver for printing Amateur and commercial Morse code and RTTY transmissions. Its small size, the built-in green video monitor screen and its 12-volt operation make the CWR6750 truly portable. The CWR6750 will receive all standard radioteleprinter speeds from 60 words per minute (45 baud) to 300 wpm (300 baud). Both the standard press "Baudot" RTTY code and the computer ASCII RTTY code can be received.

Stations using Morse code can be received at speeds from 4 to 50 wpm. A computer-style ASCII printer may be connected to the CWR6750 to provide a full printed copy of all received text.

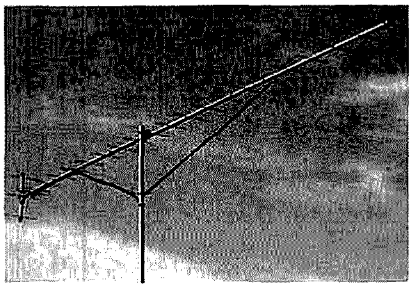
The CWR6750 measures only 10¼ × 6½ × 11 inches, and weighs only 9 pounds. It operates from any 11 to 14.5 Vdc source, drawing 1.6 amperes. The CWR6750 is easily installed in a camper, boat, or home station.



For more information, contact Hal Communications Corp., Box 365, Urbana, Illinois 61801; Reader Service Number 305.

Boomer antenna

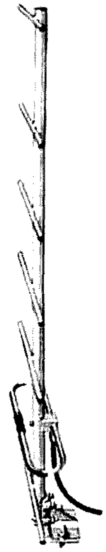
The 424B is the newest Cushcraft Boomer antenna. It is a twenty-four element, 70 cm Yagi, exhibiting 18.2 dB forward gain. A 424B



NEW products

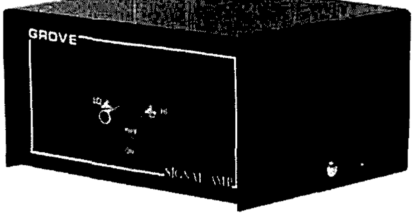
won the 70 cm antenna gain measuring contest at the 1982 Central States VHF Conference. The antenna's features include insulated elements, stainless steel hardware, N-type connector, T-match feed and trigon reflector.

For more Boomer information, contact Cushcraft Corporation, P.O. Box 4680, Manchester, New Hampshire 03108; Reader Service Number 306.



wideband antenna preamplifier

The PRE-1 Signal Amp masthead preamplifier is designed to provide high gain, low-noise amplification for received VHF and UHF signals. The PRE-1 has a midband gain of at least 15 dB with a noise figure of only 1.8 dB. The Signal Amp consists of a lightweight antenna-mounted preamplifier module and an indoor control unit. Switch-selectable high and low gain allows the user to customize his signal enhancing needs.



Guaranteed to outperform competitive indoor preamplifiers, the PRE-1 Signal Amp comes with all necessary hardware, connectors, and instruction. PRE-1 costs only $69.00 plus $2.00 UPS shipping, from Grove Enterprises, 140 Dog Branch Road, Brasstown, North Carolina 28902; Reader Service Number 307.

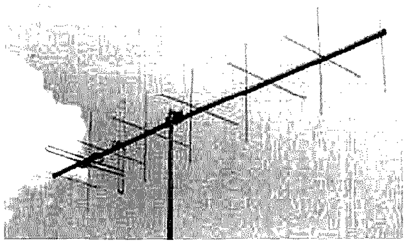
heavy-duty SRL-307 UHF Yagi antenna

Sinclair Radio Laboratories' rugged seven-element 10 dBd gain antenna will shrug off 113 mph (181 km/h) winds while carrying a 1/2-inch radial ice load, or 187 mph (301 km/h) winds without ice. This unit is useful for point-to-point links or for repeater applications in highway mobile radio systems. Reflector and director elements are 3/8-inch diameter aluminum rods welded to the boom, reducing the risk of damage and misalignment. The antenna clamp allows easy orientation for either vertical or horizontal polarization. A higher gain can be achieved by using dual (SRL-307-2) or quad (SRL-307-4) arrays with gains of 12.5 dBd and 15 dBd respectively.

For further information, contact Mr. Dan Roszelle, Sales Manager, Sinclair Radio Laboratories Inc., 14614 Grover Street, Suite 210, Omaha, Nebraska 68144; Reader Service Number 308.

circular satellite technology

The new KLM 143-150-14C circularly polarized antenna not only provides optimum reception of OSCAR satellite signals but can also dramatically improve 2-meter terrestrial communications. Linearly polarized signals (any mode, fixed or mobile) are frequently affected by buildings, mountains, and movement and, as a result, circular wavefronts develop. Reception with the 14C reduces flutter, fading, and multipath distortion, and often improves



S/N ratios. Benefits of circular polarity on transmit are similar, regardless of the polarization of the receiving antenna.

Since circularity may have a right-hand or left-hand twist, the 14C antenna kit includes a feedpoint mounted switcher, keyed by +9 to +15 Vdc. For a single-feedline convenience, a special matching harness is included. If desired, the 14C can also function as two separately fed antennas, one vertical and one horizontal. Each set of feedpoints is equipped with a 2-kW balun ready for direct coax feed.

With seven elements in each plane, the 14C produces 11-dB gain at better than 1.5:1 VSWR. Circularity is maintained within 3 dB. Virtually unbreakable 3/16 inch rod parasitic elements, anchored through the 1-1/2 inch boom, help reduce weight to 7-1/2 pounds, and windload to 1.2 square feet.

For more information, contact KLM Electronics, Inc., P.O. Box 816, Morgan Hill, California 95037; Reader Service Number 309.

R-2000 communications receiver

Trio-Kenwood has just introduced the R-2000, a highly sophisticated, all-mode communications receiver that covers 150 kHz-30 MHz in thirty bands. Designed to answer the needs of the short-wave listener as well as the Radio Amateur, this new radio is capable of receiving signals on a-m, USB, LSB, CW, and fm. Among the more interesting features to be found on this model are digital VFOs, ten memories that store frequency, band, and mode data, memory scan, programmable band scan, and dual 24-hour quartz clocks with a timer that can be programmed to turn the radio on and off on a pre-selected schedule.

Additional features include a built-in lithium battery memory back-up (estimated 5-year life), fluorescent tube digital display, three built-in i-f filters with switch, manual UP/DOWN band scan, squelch, S-meter, noise blanker, and rf step attenuator. The R-2000 operates on 100/120/220/240 Vac or it may be operated on 13.8 Vdc using an optional DCK-1 cable kit. Suggested retail price is $599.95.

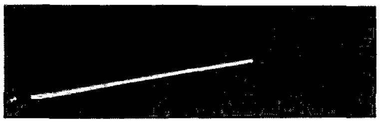
For additional information, write Trio-Kenwood Communications, 1111 West Walnut Street, Compton, California 90220. Reader Service Number 310.

NEW products

vertical mobile antennas

dB-Gain Antennas announces its new line of antenna products with the introduction of its dB-Gain vertical mobile antenna.

The antennas are available for 450 MHz, 220 MHz, 2, 6, 10, 15, 20, and 40 meters with a power rating of 250 watts. Although these antennas were designed primarily for mobile use, they can be used for a fixed station.

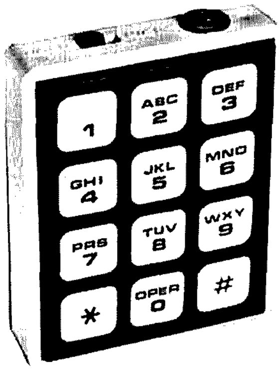


Antenna whips and set screws are made of 17-7 stainless steel for longer life and extra protection. Heavy-gauge fiberglass coil housings (0.031 wall/spiral finish) add extra strength and durability in extreme weather conditions. Each coil is wound with No. 16 copper and remaining hardware is chrome-plated brass. A standard mounting ferrule is compatible with most mobile mounts.

For more information, contact Tom Adams, W4MTW, dB-Gain Antennas, 2308 NE 20th Avenue, Ft. Lauderdale, Florida 33305; Reader Service Number 311.

smallest manual encoder

The Model 340 Thin-Coder, by CES, measures 2-1/2 × 3-1/8 × 3/4 inches. It effectively dials the user into private networks, computer access, or dimension systems. Its rugged white case features a brown faceplate and white digit blocks. A convenient normal/high switch allows flexible volume control. Up to 10,000 long distance calls are possible with the Thin-Coder's long-life 9-volt battery. CES encoders use single-contact tactile keyboards for extra reliability.



For more information on the Thin-Coder Model 340 Encoder, contact Ron Hankins, CES, Inc., P.O. Box 507, Winter Park, Florida 32790; Reader Service Number 312.

160-10 m transceiver

The FT-980 is a full-featured 160-10 meter transceiver which includes a general-coverage receiver section. Providing a nominal 100-watts rf output from a low-distortion, high-voltage final amplifier, the FT-980 is set up for full QSK with silent solid-state switching. The receiver section is designed for wide dynamic range and versatility in filter selection. An audio peak filter, i-f notch filter, variable pulse width noise blanker, variable i-f bandwidth with i-f shift (passband tuning), and an audio shaping control round out the receiver features.

The FT-980 is controlled by an 8-bit microprocessor, which allows storage of frequency and mode in memory, and programming sub-

band limits for Novice, Technician, General, or Advanced Class operators. Direct keyboard entry of frequencies provides instant QSY without the need to rotate the main tuning dial.

For more information, contact Yaesu Electronics Corp., P.O. Box 49, Paramount, California 90723. Reader Service Number 313.

indoor antenna

Contemporary Electronic Products announces the new NXL-1000 indoor shortwave antenna. Unlike other active indoor antennas, the NXL-1000 employs a Faraday shield for maximum rejection of manmade noise, so often a problem. In addition, the NXL-1000 has a built-in crystal calibrator with selectable 1-MHz and 100-kHz markers. This is a great help with uncalibrated or poorly calibrated receivers.

The NXL-1000 covers the range 1.5 through 30 MHz in three ranges. A high-Q selective circuit provides excellent rejection of unwanted frequencies, valuable for receivers with poor front-end selectivity or marginal image rejection. Internally generated noise, a problem with some active antennas, has been substantially reduced in the NXL-1000.

NEW products

the antenna is made from heavy-duty expanded aluminum screen or optional solid aluminum panels.

Named "Next Generation," this model is constructed with aircraft-style riveted aluminum framework and a single steel support for strength and light weight. It comes with a fully illustrated, step-by-step installation manual. Compatible with all popular brands of supporting electronics, the antenna is also available in colors to match the predominant local background.

For more information, contact Total Television, 17537 N. Umpqua Highway, Roseburg, Oregon 97470. Reader Service Number 315.

Eurocard racks

Designed specially for the growing interest in Eurocard-based systems, a new high-capacity rack allows placement of both single- and double-size VME-bus compatible boards in the same enclosure. The Model CCKE2, from Vector Electronic Company, also has abundant space in the rear for mounting large power supplies.

The VME bus was developed by Motorola, Mostek, Signetics, and its parent, N.V. Philips, to provide a combined sixteen-bit and thirty-two-bit standard. It employs the Eurocard format of 6.30 inches by 3.94 inches (160 mm by 100 mm) for the single card and 6.30 inches by 9.19 inches (160 mm by 233.4 mm) for the double card. Bus interconnections are made with one ninety-six-position connector on the single card and two connectors on the double card.

The CCKE2 takes advantage of the 1.3-inch (33.4-mm) difference between two single-size Eurocards and one double-size card. A simple fixture places groups of single boards one on top of another, adjacent to double boards. Appropriate system partitioning permits access to signals on either of the two VME-bus connectors.

The 19-inch (482.6-mm) EIA Std. cage holds up to twenty-seven double-size Eurocards or up to fifty-four single-size cards on 0.6-inch (15.24-mm) centers. Alternatively, the CCKE2 may be configured as a combination of Eurocard sizes; twenty-six single and thirteen double, for example.

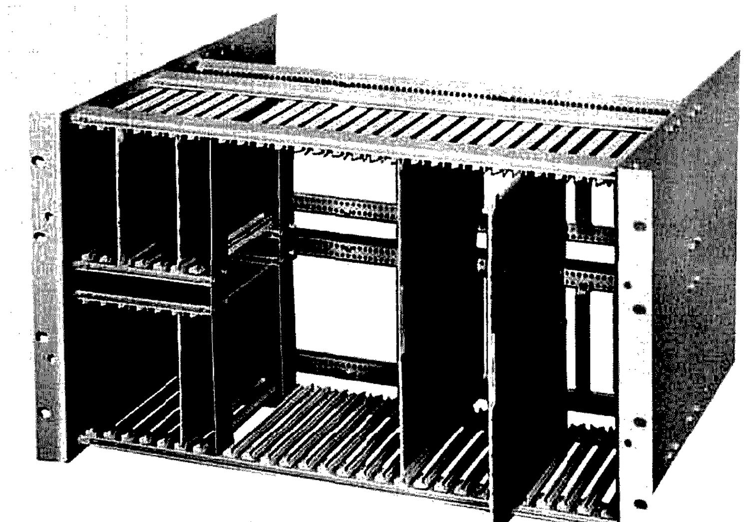
Card-guide and connector-mounting holes are spaced on 0.20-inch (5.08-mm) centers, so cards with varying component and lead heights may be installed in any position. Snap-in card guides are made of Underwriter-Laboratories-rated flame retardant grey nylon. Connectors are mounted on the pre-drilled struts with 3-48 machine screws and nuts.

At the rear of the rack, a space 10.5 inches by 16.8 inches by 5.5 inches (259.1 mm by 426.7 mm by 140 mm) is available for power supplies with 1-inch (25.4-mm) clearance for backplane wiring.

In single quantities, the fully assembled CCKE2 is priced at $68.18 each. An unassembled version, CCKE2U, is priced at $56.82 each. For more information, contact Vector Electronic Company, 12460 Gladstone Avenue, Sylmar, California 91342. Reader Service Number 316.

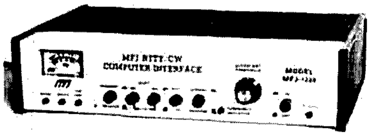
RTTY/CW computer interface

The new MFJ-1220 RTTY/CW computer interface is a terminal unit that provides TTL/CMOS and RS-232 levels for computer inter-



Universal Eurocard Racks accommodate both single and double size VME boards.

facing. Unlike phase-lock loop demodulators, this is an optimum design using individually tuned active bandpass filters. It has separate



mark and space channel filters, CW filter, and post detection lowpass filter for excellent weak-signal and high-interface RTTY/CW performance.

The MFJ-1220 takes received RTTY/CW audio from your transceiver, demodulates it, and provides TTL/CMOS and RS-232 levels for interfacing with nearly any computer. A program (not included) is used to provide RTTY/CW text.

For RTTY transmission, your computer drives the AFSK generator to provide FSK transmission using the microphone or phone patch input of your SSB transmitter, or it can directly key the FSK input of your transmitter. For CW transmission, your computer drives the high-voltage keying currents of the MFJ-1220, which then provides grid block or direct keying for your transmitter.

The RTTY/CW interface transmits and receives all standard RTTY shifts of 170, 425, and 850 Hz to cover all Amateur, commercial, and military traffic to over 100 WPM. It uses the standard space tone of 2125 Hz and marks tones of 2295, 2250, and 2975 Hz.

The MFJ-1220 RTTY/CW Computer Interface is available from MFJ Enterprises, Inc., for \$179.95 plus \$4.00 for shipping and handling. For more information, contact MFJ Enterprises, Inc., P.O. Box 494, Mississippi State, Mississippi 39762. Reader Service Number 317.

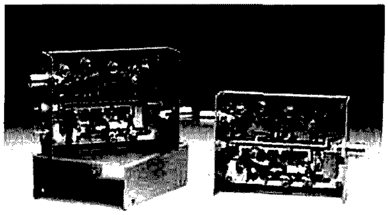
receiving converter kits

Lunar Electronics announces a new line of high-performance receiving converter kits. The initial line-up of available kits includes crystal controlled models for VHF frequencies and ultra-stable tunable oscillator models for UHF. The crystal-controlled UHF models are due out in the spring of 1983.

Easy-to-read illustrated instructions with each kit ensure the builder will achieve maximum performance from his unit. Complete factory back-up assistance, if needed, is also available. Typical specs for complete unit: input frequency 144 MHz; crystal frequency 144 MHz; image rejection −65 dB; noise figure (tune max. gain 18 dB); LO specs +7-10 dBm output; output frequency 28 MHz; conversion gain 15 dB; noise figure (tune min. NF 1.75 dB) 2.4 dB; and harmonics −50 dBc.

The highest quality components are used throughout, including double-sided, plated-through-hole PCB, gold alodined box for greatest circuit integrity, provisions for crystal netting, DBM for best performance.

NEW products



For more information, contact Lunar Electronics, 2775 Kurtz Street, Suite 11, San Diego, California 92110; Reader Service Number 318.

two antenna tuners

Encomm, Inc., announces two antenna tuners from Tokyo Hy-Power Labs, the 2000 watt HC-2000 and the 200 watt HC-200. The HC-2000 is a 2000 watt PEP (500 watts max on 1.9 MHz) hf antenna coupler with a power/SWR meter and a versatile twelve-position antenna switch (six through the tuner and six bypass). It will tune coaxial fed antennas, balanced line antennas (balun included), or end fed wires. It is band switched for 1.9, 3.5, 7, 10, 14, 18, 21, 24.5 and 28 MHz (all WARC) bands, so you don't have to experiment to find your inductor setting, plus it has 6:1 vernier dials on the capacitors for easy fine tuning. Scales on the dual meters include SWR, 2 kW, 200W, and 20W. Connectors are SO-239s and Johnson terminals. Suggested retail for the HC-2000 is $329.95.

The HC-200 is a combined 200-watt hf antenna coupler with a power/SWR meter and a six-position antenna switch (three coaxial/wire positions through the tuner and three bypass). It will tune end-fed wires, coax, or balanced line antennas (with optional balun). The HC-200 is band switched for 3.5, 7, 10, 14, 18, 21, 24.5, and 28 MHz (includes new WARC) bands. Scales on the meter include SWR, 20W, and 200W. Connectors are SO-239s and Johnson terminals. Suggested retail for the HC-200 is $99.95.

Both antenna tuners have high-quality ceramic coil forms, well damped and shielded meter circuits, as well as first-class design and layout. There are no ferrite cores in the main inductor to saturate!

For more information, contact THL Sales Department, Encomm, Inc., 2000 Ave. G, Suite 800, Plano, Texas 75074. Reader Service Number 319.

80-MHz multifunction counter

A new 80-MHz, eight-digit multifunction counter that provides frequency, period, and totalize measurements has been introduced by the B&K Precision Test Instrument Product Group of Dynascan Corporation. Designated Model 1805, this lightweight unit measures frequencies from 5 Hz to 80 MHz. Resolution may be selected from 0.1 Hz for frequencies below 10 MHz to 1 Hz for frequencies above 10 MHz. The period mode can be used to measure low frequencies from 5 Hz to 2 MHz more accurately. The totalize mode counts individual events from 0 to 99,999,999 with an overflow LED. This model is helpful in applications where a specific number of cycles occurs, such as gated tone bursts.

The B&K-Precision Model 1805 utilizes a 10-MHz time base generated by a crystal controlled oscillator for good stability with regard to temperature (<0.001 percent ± 10 ppm at 0 degrees C - 50 degrees) and line voltage variations (< ± 1 ppm with ± 10 percent line voltage regulation). For lessened susceptibility to noise and undesirable high-frequency components, a front-panel-switchable 100-kHz low-pass filter is incorporated in the counter. All operating modes, resolution ranges, and functions are front-panel selectable. The Model 1805 incorporates a switchable X10 attenuator, **HOLD** switch to freeze the display at the present reading, and a **RESET** switch to clear the display and initiate a new measurement.

The Model 1805 is available from B&K-Precision Electronic distributors. Suggested price is $290.00. For further information, contact B&K-Precision Test Instrument Product Group, Dynascan Corporation, 6460 W. Cortland Street, Chicago, Illinois 60635. Reader Service Number 320.

Ham software available. Information: AnTech, POB 8964, Fort Collins, CO 80525.

MOBILE IGNITION SHIELDING. Estes Engineering, 930 Marine Dr., Port Angeles, WA 98362.

CABLE CONVERTERS, decoders. Free catalog. APS, POB 263 HR, Newport, RI 02840.

RTTY EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. PO Box RY, Cardiff, CA 92007.

TELEGRAPH AND WIRELESS keys wanted. Advise of identification, condition and postpaid price. Dick Randall, K6ARE, 1263 Lakehurst, Livermore, CA 94550.

MANUALS for most ham gear made 1937/1970. Send $1.00 for 18 page "Manual List", postpaid. HI-MANUALS, Box R802, Council Bluffs, Iowa 51502.

SATELLITE TELEVISION: Free wholesale price list. Know the facts with "Handbook and Buyers Guide" only $9.95. Communications Consultants, PO Box 5099, Fort Smith, Arkansas 72913.

WANTED: Schematics-Rider, Sams or other early publications. Scaramella, P.O. Box 1, Woonsocket, RI 02895-0001.

RETIRING to Florida. Must sell 30-year accumulation of Ham gear, parts, Ham and technical books immediately. Sacrifice prices. Send SASE for list. Dino Mastrojohn, 10 Madeleine Pl., Parsippany, NJ 07054. W2UII.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

RUBBER STAMPS: 3 lines $3.25 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

WANTED: New or used MS and coaxial connectors, synchros, tubes, components, military surplus equipment. Bill Williams, PO #7057, Norfolk, VA 23509.

VERY in-ter-est-ing! Next 5 issues $2. Ham Trader "Yellow Sheets", POB356, Wheaton, IL 60189.

CB TO 10 METER PROFESSIONALS: Your rig or buy ours — AM/FM/SSB/CW. Certified Communications, 4138 So. Ferris, Fremont, Michigan 49412; (616) 924-4561.

HAMS FOR CHRIST — Reach other Hams with a Gospel Tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

PLANS, CIRCUIT BOARDS, AND KIT PARTS (author approved) for Leach's construction projects. Power amps, preamps, pre-amps and loudspeakers. Send SASE for information. Custom Components, Box 33193, Decatur, GA 30033.

Coming Events
ACTIVITIES
"Places to go..."

ARKANSAS: The Northwest Arkansas Amateur Radio Club's 3rd annual Hamfest/Swapmeet, Saturday, May 21, Rogers Youth Center, 315 West Olive Street, Rogers, 8 AM to 4 PM. Free admission. Commercial exhibitors and flea market tables/space $2.00. Setup 6 AM. Free parking. Refreshments nearby. Snack bar on premises. Talk in on 146.16/.76 or 146.52 simplex. For more information: Mary Webb, KA5HEV, PO Box 338, Prairie Grove, AR 72753.

CALIFORNIA: The North Hills Radio Club's 11th annual Sacramento Valley Hamswap, May 1, 9 AM to 3 PM, Placer County Fairgrounds, Roseville. Free admission. Tables $6 to $8. Tailgate sites $5.00. Talk in on K6IS repeater (144.59/145.19). For information: Doug Long, KB6ZR, 8810 Swallow Way, Fair Oaks, CA 95628. (916) 961-0728.

CALIFORNIA: West Coast VHF/UHF Conference sponsored by W6GD UHF Society, May 7 and 8, Sunnyvale Hilton Inn, 1250 Lakeside Drive, Sunnyvale. $8 pre-registration by April 27, 1983. $10 door. Displays, programs, DX and contest operating, computers, swap and flea market. Saturday evening banquet. For information: West Coast VHF/UHF Conference, PO Box 4101, Fremont, CA 94539.

COLORADO: The Rocky Mountain VHF Society's annual Swapfest, Sunday, May 22, 9 AM to 4 PM, Colorado National Guard Armory, 4750 North Broadway, Boulder.

Admission $2.00. No additional charge for display tables. Seminars, refreshments and lots of fun! Talk in on 146.16/76 and 146.52 simplex. For additinal information call (303) 494-6291.

FLORIDA: When visiting Florida, stop in at a Welcome Center and get a complimentary 2-meter directory and a refreshing' glass of juice. *Cloverleaf Farms Amateur* Radio Club and Cloverleaf Farms Manufactured Homes Community have cooperated to bring this service to hams traveling in Florida.

GEORGIA: The Anderson, Hartwell and Toccoa Amateur Radio Clubs' 5th annual Lake Hartwell Hamfest, May 21 and 22, Lake Hartwell Group Camp, Hartwell. Free admission, free camping and free flea market space. A left-footed CW contest, horseshoes, fishing, swimming and more for the whole family. Campgrounds open 6 PM Friday evening. Talk in on 146.19/.79, 147.93/.33 and 146.895/.295. For further information: Ray Pettit, WB4ZLG, Rt. #1, Dooley Drive, Toccoa, GA 30577.

IDAHO: Kootenai Amateur Radio Society's Hamfest '83, Saturday, June 11, North Idaho Fairgrounds, Coeur d'Alene, 8 AM to 4 PM. Free swap tables, large RV parking area. Food available. Talk in on 146.38/98 or 146.52 simplex. For further information: Vladimir J. Kalina, South 1555 Signal Point Road, Post Falls, ID 83854.

ILLINOIS: The Six Meter Club of Chicago is having their 26th annual Hamfest, Sunday, June 12, Santa Fe Park, 91st and Wolf Road, Willow Springs, southwest of Chicago. Gates open 6 AM. Advance registration $2.00, $3.00 at gate. Large swapper's row, picnicking, pavilion displays, refreshments, AFMARS Meeting. Talk in on K9ONA 146.52 or K9ONA/R 37-97. For advance tickets: Val Hellwig, K9ZWV, 3420 South 60th Court, Cicero, IL 60650.

INDIANA: The 4th annual MAARC Hamfest, May 22, Delaware County Fairgrounds, 8 AM to 3 PM. All activities inside. Fla market tables $5.00. Tickets $2.00 advance, $3.00 at door. Free parking. Food, forums, computer displays. Talk in on 146.13/73, 146.52, 223.10/224.70. For further information: Craig Graham, WD9EHF, RR 12, Box 86, Muncie, IN 47302.

INDIANA: The Wabash Valley Amateur Radio Association's 37th annual Hamfest, Sunday, June 5, Vigo County Fairgrounds, Terre Haute. For more information SASE to W.V.A.R.A., PO Box 81, Terre Haute, IN 47808.

INDIANA: The Tristate Amateur Radio Society's annual Hamfest, Sunday, May 15, Vanderburgh County 4H Center, Evansville. Admission $2. Open 6 AM CDT. Indoor tables available. Outdoor flea market. Talk in on 147.75/.15 and 146.19/.79. For information and table reservations: Hal Wilson, WB9FNN, RR #8, Box 427B, Evansville, IN 47711.

KANSAS: The Central Kansas Amateur Radio Club's 3rd annual Kansas State ARRL Convention, June 4 and 5, Red Coach Inn Convention Center, West Crawford and I-135, Salina. Programs for Hams, non-Hams and ladies. Free flea market adjacent to Center. Saturday evening banquet and entertainment. For further information SASE to Bill Ringquist, KAØCUF, RR #1 Box 155, Gypsum, KS 67448.

KANSAS: The Pittsburg Repeater Organization's annual Hamfest, May 15, 10 AM to 5 PM, Lincoln Center, Lincoln Park, Pittsburg. Covered dish dinner, flea market. Admission $1.00 at door.

KENTUCKY: Northern Kentucky Amateur Radio Club's annual Ham-A-Rama, Sunday, June 5, Burlington Fairgrounds, Burlington. Tickets $5.00 at gate. Flea market space $3.00. Vendors, nets and group meetings. Refreshments available. Talk in on 147/86 and 375/975. For information: Dick Johnston, WA4KUB, 3113 Brookwood Dr., Edgewood, KY 41017. (606) 341-8759.

MARYLAND: The Maryland FM Association's annual Hamfest, Sunday, May 29, Howard County Fairgrounds, West Friendship. 8 AM to 4 PM. Donation $3.00. Tailgating $3.00. Inside tables in advance $6.00 each, at door $10.00 each. Talk in on 146.16/76 and 146.52. For information and table reservations: MFMA HAMFEST COMMITTEE, c/o John Elgin, WA3MNN, 5495 Apt 2, Harpers Farm Road, Columbia, MD 21044. (301) 596-3741.

MICHIGAN: The Chelsea Communications Club is sponsoring a Swap 'N Shop, Sunday, June 5, Chelsea Fairgrounds. 8 AM to 2 PM. Gates open for sellers 5 AM. Donation $2.50 advance, $3.00 door. Children under 12 and non-ham spouses admitted free. Talk in on 146.52 simplex and 147.855 Chelsea repeater. For information: William Altenberndt, 3132 Timberline, Jackson, MI 49201.

MICHIGAN: The Independent Repeater Association of Grand Rapids will hold its annual *Hamfestival, Saturday*, June 4, 8 AM to 4 PM, Wyoming National Guard Armory on 44th St. east of US-131. Dealer setup 6 AM. Free table space to all sellers. Admission $3.50. ATV, satellites, contests, computers, MARS and schack photo contest. Huge swap area. Talk in on 147.165/147.765. For information and table reservations: John Knoper, KC8KK, (616)

534-5501 or I.R.A., 562 - 92nd Street S.E., Byron Center, MI 49315.

MINNESOTA: NARA is again sponsoring the state's largest swapfest and exposition of personal computers and software, June 4, Minnesota State Fairgrounds, Snelling Avenue, north of I-94. Large indoor commercial exhibits and booths. Giant outdoor flea market. Admission $4.00. For more information or dealer inquiries: Amateur Fair, PO Box 857, Hopkins, MN 55343 (612) 420-6000.

MISSOURI: The Indian Foothills Amateur Radio Club's 8th annual Hamfest, Sunday, May 15, Saline County Fairgrounds, Marshall. Tickets $2.00 each, 3/$5.00 at door, 4/$5.00 advance. Registration 8 AM. Free flea market tables, registration required. Talk in on 52, 147.84/.24. For information and tickets: Fred Fellers, WØABW, 703 N. Main, Carrollton, MO 64633. (816) 542-0223 or 542-2655 or (816) 886-2837.

NEW ENGLAND: The Hosstraders will hold their tenth annual Tailgate Swapfest, Saturday, May 7, sunrise to sunset, at Deerfield, NH, Fairgrounds. Admission $1.00, including tailgaters and commercial. Friday night camping for self-contained rigs at nominal fee. None admitted before 4 PM Friday. Profits benefit Boston Burns Unit of Shriners' Hospital. Last year's donation $2622.75. Questions or map to northeast's biggest ham flea market? SASE to Norm, WA1IVB, RFD Box 57, West Baldwin, ME 04091 or Joe, K1RQG, Star Route, Box 56, Bucksport, ME 04416 or Bob, W1GWU, North Walton Road, Seabrook, NH.

NEW HAMPSHIRE: The 9th annual Eastern VHF/UHF Conference, May 13-15, Sheraton Tara, Nashua. Friday night hospitality room. Saturday night banquet, $14, payable prior to May 9. Registration $13.50 from K1LOG, Rick Commo, 3 Pryor Rd., Natick, MA 01760 before May 9. Registration at door $20.00.

NEW JERSEY: The Jersey Shore Chaverim are sponsoring the Jersey Shore Hamfest and electronic flea market, June 12, 9 AM to 3:30 PM, Jewish Community Center, 100 Grand Avenue, Deal. Admission $3 per person. Children under 12 and XYLs free. Refreshments available. Table $5. Tailgating $2.50. Reserve spaces by SASE and advance payment to Jersey Shore Hamfest, PO Box 192, West Long Branch, NJ 07764 by May 15. Talk in on 147.045 + .6; 146.52 simplex.

NEW YORK: The Rochester Hamfest combined with ARRL New York State and Atlantic Division Conventions, May 20 and 21, Marriott Thruway Hotel and Monroe County Fairgrounds. Tickets $4 advance and $5 at gate. Flea market tickets $2 per space. FCC exams given. Send Form 610 to FCC, 1307 Federal Building, 111 W. Huron St., Buffalo, NY 14202 by May 1 marked "administered at Rochester Hamfest." Friday evening banquet (instead of Saturday). Flea market open 6 AM Saturday, commercial exhibits 8:30 AM. Closing time 6:00 PM. Talk in on 146.28/88 and 144.51/145.11. Advance tickets from K2MP, 737 Latta Road, Rochester, NY 14612. For more information: Rochester Hamfest, 300 White Spruce Blvd., Rochester, NY 14623.

NEW YORK: The Putnam Emergency Amateur Repeater League (PEARL) will have its 2nd annual indoor Hamfest, Saturday, May 7, 9 AM to 4 PM, JFK Elementary School, Foggintown Road, Brewster. General admission $1.00. Exhibitors $4.00. Talk in on 144.535/145.135 and 52. For advance table registration and information: Frank Konecnik, WB2PTP, RD 1 - 244 C, Carmel, NY 10512.

NEW YORK: The 24th annual Southern Tier Amateur Radio Club's Hamfest, Saturday, May 7, Treadway Inn, Owego. Flea market opens at 8 AM. Vendor displays and sales. Tech and non-tech talks. Refreshments. Advance tickets only for the dinner at 6:30. Talk in on 22/82, 16/76 or 146.52 simplex. For further information SASE to KF2X, C. England, RD #1, Box 144, Vestal, NY 13850.

NEW YORK: The Ebonaire Amateur Radio Society's 2nd annual Hamfest/Flea Market, Sunday, June 5, 9 AM to 3 PM, 119-09 Merrick Blvd., Queens. Contact WA2VYG (212) 523-2319 or KA2CPA (212) 528-0416.

NEW YORK: The Rome Radio Club's 31st Rome Ham Family Day, Sunday, June 5, Beck's Grove in Rome. Games, contests, technical presentations and a giant flea market are some of the features. Refreshments available throughout the day. The Club's "Ham of the Year" award will be presented at the buffet dinner. Talk in on 146.28/88 and 146.52 simplex.

NORTH CAROLINA: Durhamfest sponsored by the Durham FM Association, Saturday, May 14, South Square Shopping Center, Durham. Flea market, dealers, tables available for rent. Admission $4.00. Talk in on 147.825/.225 and 146.52 simplex. For information: DFMA, PO Box 8651, Durham, NC 27707.

OHIO: The Fremont Radio Club in cooperation with the Ottawa County Radio Club is sponsoring their 6th annual Hamfest, May 22, Fremont Fairgrounds. Gates open 8 AM. Dealer setup 7 AM. Advance tickets $2.50. $3.00 at door. Flea market tables $3.00 per 8 ft. space. For tickets and table reservations SASE to John Dickey, W8CDR, 545 N. Jackson Street, Fremont, OH 43420. (419) 332-8066.

OHIO: The Lancaster and Fairfield County Amateur

Radio Club's annual Fathers' Day Hamfest, Sunday, June 19, Fairfield County Fairgrounds, Lancaster, 8 AM to 4 PM. Admission $2.00 advance; $3.00 at gate. Refreshments available. Free parking. Many covered tables. Talk in on 147.03/63 or 146.52 simplex. For more information: Box #3, Lancaster, OH 43130.

PENNSYLVANIA: The 29th annual Breeze Shooters Hamfest, Sunday, May 22, 9 AM to 5 PM, White Swan Amusement Park, PA, Rt. 60 near the Greater Pittsburgh International Airport. Free admission. Free flea market. Registration $2.00 or 3/$5.00. Covered vendors tables by advance registration. Talk in on 146.28/.88 or 29.0 MHz. Contact Don Myslewski, K3CHD, 359 McMahon Road, North Huntingdon, PA 15642. (412)863-0570.

PENNSYLVANIA: The Warminster Amateur Radio Club's annual Hamfest, Sunday, May 15, Middletown Grange Fairgrounds, Penns Park Road, Wrightstown (Phila. area) 7 AM to 2 PM. Admission $3.00 per ham. Sellers $2.00 additional per 8 ft. space. Inside spaces available. No power. Registration prior to May 1, $2.00 per ham. Talk in on 147.69/09 and 146.52 simplex. For information: WARC, Box 113, Warminster, PA 18974. Or call Frank, AK3O (215) 968-3133 after 2300 UTC.

ROCHESTER HAMFEST: Atlantic Division/New York State Convention. Saturday, May 21, Monroe County Fairgrounds. Hotel headquarters, Rochester Marriott Thruway. More info? Write or call Rochester Hamfest, 300 White Spruce Blvd., Rochester, NY 14623 (716) 424-7184.

SOUTH CAROLINA: The Blue Ridge Amateur Radio Society's Hamfest, Saturday, April 30 and Sunday, May 1, at the American Legion Fairgrounds, White Horse Road, Greenville. Admission $3.00. Talk in on 146.01/61 and 223.46/224.06. For information: Phil Mullins, WD4KTG, Hamfest Chairman, PO Box 99, Simpsonville, SC 29681. For advance sales: Mrs. Sue Chism, Rt. 6, 203 Lanewood Dr., Greenville, SC 29607.

TENNESSEE: The Radio Amateur Club of Knox County will hold its 17th annual Hamfest, Saturday, May 28, 9-5 and Sunday May 29, 10-4, Kerbella Temple Auditorium, east of US 441 behind Vol Inn Motel. Admission $2.00 advance, $3.00 at door. Radio and computer forums, dealers, indoor and tailgate flea markets. Free parking. Talk in on 147.90/30. For tickets, dealer or flea market information: Mark Nelson, AJ2X, 4317 Foley Drive, Knoxville, TN 37918. (615) 687-9656.

TEXAS: The YL International Single Sidebander's 1983 Convention, June 16-19, Dallas. Activities include the DX Roundup, the System Awards banquet Saturday night with a country-western band for dancing. Preconvention activities begin June 13. For detailed information: Joe, W5UJO and Mary, KC5UO, Parsons, 1639 Evergreen Drive, Mesquite, TX 75149.

VIRGINIA: Mayfest '83 presented by the Roanoke Valley Amateur Radio Club, Sunday, May 29, 0900 to 1600, Roanoke Civic Center Exhibit Hall. Advance registration $3.00, $3.50 at door. CW contest, ARRL forum YL, XYL and kiddie functions. Nearby motels, camping and sightseeing. Talk in on 146.385/.985 and 146.52 simplex. For information, tickets and tables: Bill Johnson, W4NLC, 5129-D Overland Rd., Roanoke, VA 24014. (703) 989-5374.

WASHINGTON: The Tri-Cities Hamfest Council 4th annual Hamfest, May 21 and 22, starting 9 AM, Benton-Franklin County Fairgrounds, Kennewick. Admission $3.00 advance, $4.00 at door. Children under 12 free. Vendors, swap tables, Bunny Hunt on Sunday morning. Camping and RV space at site $6.00. For reservations and information: (509) 586-9375 or (509) 967-2358. Inquiries to Tri-City Hamfest Council, PO Box 1181, Richland, WA 99352.

OPERATING EVENTS

"Things to do..."

MAY 14: Ling Submarine Expedition. The Meadowlands Amateur Radio Association will be aboard the USS Ling (SS297) docked in Hackensack, New Jersey, and will operate under club station N2BMN, starting Saturday at 1500Z through 2100Z. 20 Meters: CW 14.060, SSB 14.310. 40 Meters: CW 7.115, SSB 7.250. 2 Meters: CW 144.100, SSB 144.160, FM 146.550. 6 Meters: CW 50.095, SSB 50.125. For an 8½ × 11 certificate to confirm OSO send large SASE with 37¢ U.S. Postage to PO Box 324, Little Ferry, NJ 07643.

MAY 16-21: Jimmy Stewart's Birthday. The Indiana (PA) County ARC will help the community of Indiana, PA, celebrate this native son's 75th birthday. Club members will be on all General and Novice frequencies at various times and frequencies. SASE with QSL card to W3FVU for a commemorative QSL card.

MAY 21: ARMED FORCES DAY military-to-Amateur cross band operations will be conducted from 21/1300 UTC to 22/0245 UTC May 1983. East coast stations commence operations at 21/1300 UTC and west coast stations commence operations at 21/1600 UTC. Military stations will transmit on selected military frequencies and listen for Amateur stations on the specific frequency to

which she/he is listening. Entries must be postmarked no later than 28 May 1983 and submitted to the respective military commands. Stations copying AIR send entries to: Armed Forces Day Test, 2045th CG/DONJM, Andrews AFB, DC 20331. Stations copying NAM, NAV or NPG send entries to: Armed Forces Day Test, HQ, Navy-Marine Corps MARS, 4401 Massachusetts Ave., N.W., Washington, DC 20390. Stations copying WAR send entries to: Armed Forces Day Test, Commander, 7th Signal Command, Att: CCN-PO-DX, Fort Ritchie, MD 21719.

MAY 21 AND 22: The Clark County Amateur Radio Club, W7AIA, is pleased to announce the third annual Mount Saint Helens QSO party to mark the third anniversary of the explosion of nearby Mt. Saint Helens. 0001 UTC May 21 through 2359 UTC May 22. Look for W7AIA on: SSB 3.895, 7.230, 14.280, 21.360, 28.505. CW 3.705, 7.105, 21.105, 28.105. VHF — various Vancouver and Portland area repeaters. To apply for the award send log information or QSL card and $2.00 (or 8 IRCs) to: Award Manager, W7AIA, PO Box 1424, Vancouver, WA 98668.

MAY 28 AND 29: The Northwest Amateur Radio Club will operate W9LM from 1700Z May 28 to 1700Z May 29 to commemorate their 50 years in Amateur Radio. Frequencies: Phone 10 kHz from lower General 40, 20, 15, and 10. CW 25 kHz from lower edge of Novice bands and 2 meter simplex on 146.52. QSL with SASE for commemorative certificate to: NARC, PO Box 121, Arlington Heights, IL 60006.

JUNE 2, 3 AND 4: S.P.A.R.C., the Southern Piedmont Amateur Radio Club, will operate a special event station, the 10th annual "Helen to the Atlantic Ocean" hot air balloon race, held under the direction of the "Free Spirits of Helen, Inc." The station will be operating SSB between 7200-7250 and 3865-3915 on 40 and 80 meters using club call WD4NHW. For an 8 × 10 certificate send SASE to John Anthony, PO Box 28, Sautee, GA 30571.

JUNE 4: The Pennyroyal Amateur Radio Society announces the annual Jefferson Davis QSO party, Saturday, 1500 to 2400 Z. Suggested frequencies: 3.940, 7.260, 14.310, 21.410 and 28.610 MHz phone and 3.730 MHz CW. For an attractive certificate send $1.00 and 3/20¢ stamps with QSL card to P.A.R.S., PO Box 1077, Hopkinsville, KY 42240.

JUNE 10 AND 12: The Wireless Institute of Northern Ohio (W.I.N.O.) will operate a special events station (KO8O) from a winery in Madison, Ohio, to commemorate Ohio Wine Week. Friday 2300Z to 0300Z on 3900 MHz and 7235 MHz. Sunday 1500Z to 2000Z on 7235 MHz and 21360 MHz. For a special QSL certificate send legal SASE with 40¢ postage or coin to above address.

JUNE 1983 / $2.50

ham radio magazine

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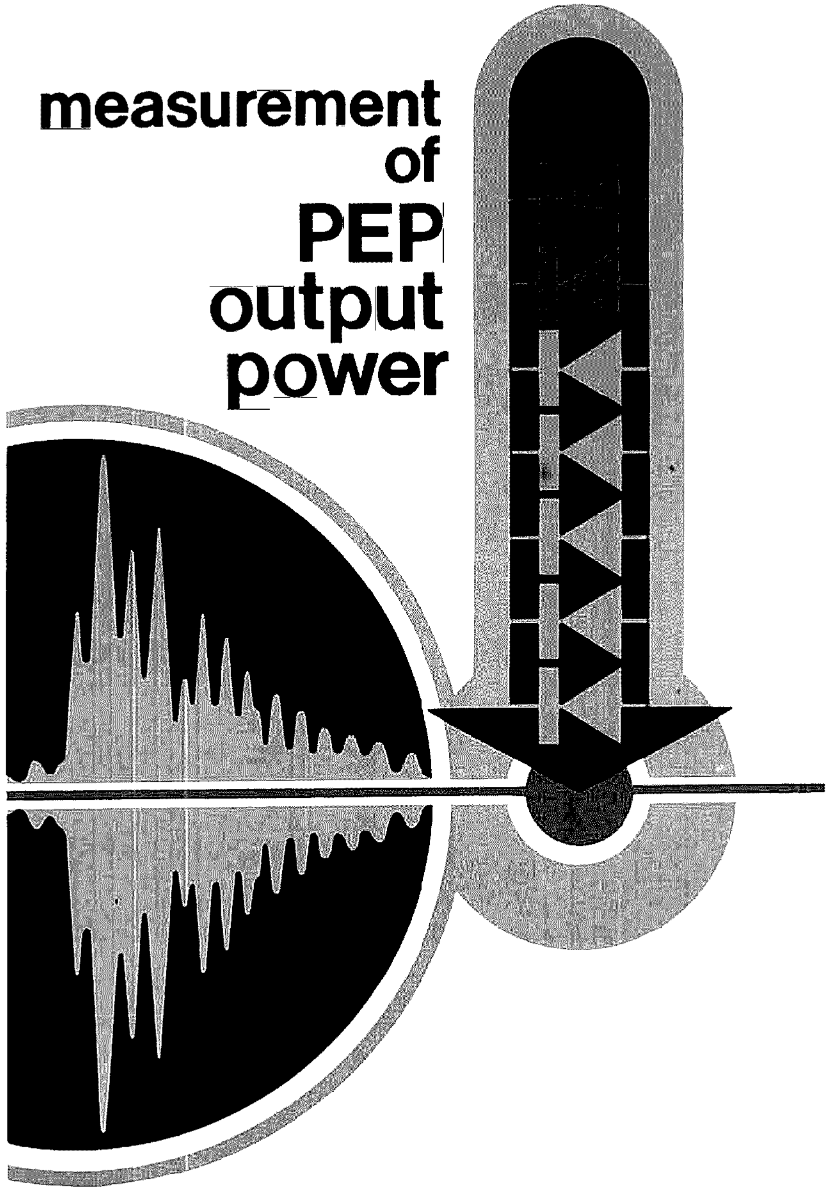

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contents

REFLECTIONS

window on the world — the "foreign" press

It's always interesting to examine the approaches taken by others in solving problems common to all of us.

At *ham radio*, we have the opportunity to provide readers with a window on the world of Amateur Radio. Dozens of foreign language publications pass over our desk each month*; and as we promised last year, we'll do all we can to provide technical notes, articles, and other items of interest carefully selected from these numerous and informative sources.

In our lead article VK3AFQ provides us with a view of technical requirements facing Australian hams in 1983. Not surprisingly, we, here in the States, have the same situation — how to specify and measure rf power.

Up till now we have met the FCC stipulated power requirements by measuring the dc power input to the final amplifier stage (including drive) while keeping it under 1 kW. Recent discussions have focused on increasing that power level to 1500 watts of PEP *output*. It's a simple enough matter to measure dc power input — plate voltage x plate current. But how do you measure PEP output? We return to the land down under and see how VK3AFQ solved the problem simply and effectively in his article, "Measurement of PEP Output Power."

While still "down under," three magazines that have provided Australian and New Zealand hams with the latest in technical information are *Amateur Radio*, (VK); *Amateur Radio Action*, (VK); and *Break-In*, (ZL). Ron Cook, VK3AFW, in his "Novice Notes" (*Amateur Radio*, February, 1983) addresses a situation many of us now face — the need to provide a low VSWR load (antenna) to our solid-state rigs. His article discusses antenna tuners including the L, T, and Pi networks, the ultimate transmatch, and a special wide-range tuner while providing theory and application information.

A very popular antenna these days, especially with the lower band (40, 80, 160) enthusiasts is the delta loop. Probably one of the best descriptions of its operation recently appeared in *Amateur Radio Action*. In the March, 1983 issue, VK2EAO details "Loaded Corner-Fed Delta Loops." After a brief historical description of its development the author provides construction details extending this antenna's use to any of the hf ham bands.

Satellite users will find some notes of interest in New Zealand's Amateur journal *Break-In*. The regularly featured column "Satellite News" by ZL1TGC provides the latest data on the Phase III-B, UO-9, RS-8, and RK-03 (Iskra-3) "birds" perhaps from a different *slant*. For example, if you'd like to know what the UoSAT satellite is experiencing in space, examine the detailed telemetry information from its sensors provided on channels 00 through 59 as delineated in this February's column.

Rotating our sights to the northwest we briefly pause to examine Japan's monthly journal, *CQ ham radio*. This 500 plus page magazine features columns for SSTV, RTTY, VHF, UHF, and Microwave enthusiasts, to name just a few. For example, in the March, 1983 issue, construction details, polar and VSWR plots are provided for a 36-element 2400 MHz single boom Yagi. Being a lower band "aficionado," I can't help but notice the 14 page DX "column" and wonder how I could possibly have recently missed so many 80-meter stations such as A92NH, DU1PJS, HLØCCA, 5V7AL and 9N1RFT — all heard from Japan. It's necessary to mention that except for a single page column called "The English Service of CQ ham radio," all the rest of the magazine is in Japanese.

Continuing in a clockwise rotation, we arrive at the British Isles and notice English language publications such as *Radio Communications*, *The Short Wave Magazine*, *Practical Wireless*, and *Radio and Electronics World*. Though the last one is not specifically dedicated to Amateur Radio, it includes several articles of direct interest to hams. An excellent article by A.J. Rogers in the April, 1983, issue on "Crystal Filter Design" provides useful data for the experimenter to enable him to develop his own. One example illustrated in the article is a 6-crystal filter featuring a half-power bandwidth of ±4.5 kHz, 60 dB bandwidth of ±12.5 kHz with a maximum passband ripple and insertion loss of 1 and 1.75 dB, respectively. Imagine putting one of these *in front* of your receiver on your favorite band. It should go a long way toward reducing that neighboring ham's strong signal 20 kHz away from you.

From this editor's position, I see so many articles that would be of interest to you, the reader. For example, across the channel in France we have two excellent ham magazines, *Megahertz* and *Ondes Courtes Informations* (Short Wave Information). This month they feature articles on . . . But wait. I'll tell you about them and others in future issues — if you'd like. Let me know your thoughts.

Rich Rosen, K2RR
Editor-in-Chief

*An SASE to me will provide you with the latest address and subscription information on these journals.

presstop de W9JUV

TRAGEDY IN THE SOUTH CHINA SEA ENDED A DXPEDITION to Spratly Island April 10. Two of the four Amateurs in the all-German group died as a consequence of Vietnamese shellfire after their chartered catamaran strayed too close to an occupied atoll in the hotly disputed islands. After the vivid on-the-air account of the attack ended abruptly, the fate of the four DXers, their vessel and its two-person crew was unknown for over a week until a Panamanian freighter, the Linden, pulled the four survivors from a liferaft.
DJ6SI and DF6FK Survived The Unprovoked Attack, while DJ3NG and DJ4EF both died.

THE ARRL'S OPPOSITION TO A "NO-CODE" LICENSE WAS FIRMLY MAINTAINED by the directors at their April 21-22 meeting. Despite recent indications of a more conciliatory attitude on the part of some League staff, and strong support for a codeless license by several of the directors, the outcome of the meeting was a resolution "stating in the strongest possible terms the opposition of the League to the creation of an Amateur license class that does not require demonstration of a knowledge of Morse Code."

ADDITIONAL PHONE FREQUENCIES ON 75, 15, AND 10 METERS were proposed in a new FCC Notice of Proposed Rule Making that accompanied the 20-meter expansion announced in last month's Presstop. On 75 meters the FCC proposes moving the lower phone limit down 25 kHz to 3750 for Extras, to 3775 for Advanced, and to 3850 for Generals. On 15 meters 21200 would be the new bottom edge for Extras, 21225 for Advanced, and 21300 for Generals. The 10 meter phone band would start at 28300, a full 200 kHz lower for these three license classes.
40 Meters Would Remain As Is For All U.S. Amateurs except those in Hawaii and nearby islands. In those areas the phone band edge would be lowered to 7075 kHz, to bring their phone privileges in line with those of others in the Pacific area.
Comments On The Additional Expansion Are Due July 1st; Reply Comments August 1. The same docket number as before, PR Docket 82-83, still applies.
The New 14150-14200 Phone Allocation Became Available to U.S. Amateurs May 22.
The Commission Proposal To Extend The Term Of Amateur Licenses to 10 years has been designated PR Docket 83-337. Comments are due at the FCC June 13; Reply Comments July 13.

LAUNCH OF THE PHASE III-B SPACECRAFT HAS BEEN DELAYED AGAIN, probably into July. After checkout of the new Amateur satellite was completed in Germany, AMSAT engineering VP W3IWI and AMSAT-DL Amateurs took the bird to the Kourou, French Guiana launch site for integration with the Ariane launch vehicle. Now, however, problems have developed with Ariane that are likely to push the actual launch back at least another month.
An Extensive Launch Day Net Operation Has Been Planned, to provide AMSAT members and other interested Amateurs with last minute developments as the launch proceeds. The launch day net will begin operation at 1130Z the day of the launch on the regular AMSAT net frequencies: 3850, 7280, 14282, and 21280 kHz. Net operation will continue through the launch window (about 3 hours) or until the launch vehicle and its cargo are in orbit.
For The New Launch Date Check The AMSAT Domestic Nets on Tuesday evening (Wednesday morning GMT) starting at 0200Z on 3850 kHz, or W1AW bulletin.

CB AND RADIO CONTROL LICENSING WAS ELIMINATED at the FCC April 27th Agenda Meeting. Effective immediately no CB or RC licenses will be processed though the effective date had not been set at press time.

W5LFL/SPACESHIP MOBILE IS NOW DEFINITE and will become a reality when the Spaceship Columbia carries the joint U.S.-European Spacelab aloft on the STS-9 flight the end of September. Operation is to be simplex on the low end of 2 meters to give Amateurs in other ITU regions, where the band is 144-146 MHz, a chance to work W5LFL. He'll operate about an hour a day, during "off duty" periods, with an operating protocol that's still being worked out that's hoped will avoid the typical DXpedition "pileups."
Operation Will Be Conducted By FCC Rules Under International Procedures established at the 1971 Space Conference. These require formal notification of the appropriate agency well in advance of such an operation, though at press time no notification had been received by the FCC. Scheduled date of the STS-9 launch is September 30.

DAYTON HAMVENTION'S "HAM OF THE YEAR" IS KH6IJ, who's internationally known as one of the world's top contest operators but whose lifelong efforts in promoting Amateur Radio are equally impressive if less well known. The Hamvention "Special Achievement Award" went to Lenore Jensen, W6NAZ, for her outstanding success in promoting public recognition of Amateur Radio. It's been principally through her efforts that so many celebrities took part in the ARRL sponsored Public Service Announcement program. Congratulations to both!

FCC COMMISSIONER ANNE JONES HAS ANNOUNCED HER RESIGNATION from the Commission to return to private law practice. Ms. Jones has been a real friend of Amateur Radio since becoming a Commissioner in 1975, with a better-than-average understanding of Amateur concerns. Her valuable contributions will be missed.

ADDITIONAL FREQUENCIES FOR CORDLESS PHONES WERE PROPOSED by the FCC in a March 31 Notice of Proposed Rule Making. The new frequencies, between 46.6 and 47 MHz, would supplement the present 49.6-50 MHz frequencies on a five-year interim basis until a more suitable permanent spot could be found for the burgeoning portable phone industry.
Interference To Amateur 160 And 80 Meter Operation by the base station portion of cordless phone systems, which operate around 1.7 MHz, has become a growing problem. It's hoped this new FCC proposal, General Docket 83-325, can help to end this interference.

comments

switching circuit

Dear HR:

The switching circuit described by Fred Dahnke, WB6IQV, on page 70 of the January, 1983, issue of *ham radio* may cause problems in many of the newer GM automobile radio systems. Neither side of the speakers in the newer systems is grounded, and introducing a ground, as in the drawing, may cause component failure within the radio.

Modifications to the circuit may be made so the speaker system remains balanced, both with the speaker, and a load resistor, yet work with the unbalanced configuration for the two-way radio. On the other hand, some consideration must be given to the fact the car radio speaker system will be open during the switching time of the relay, and the possibility of damage to the radio because of this.

The final decision is up to the vehicle owner, and the above must be weighed in this consideration.

Sheldon Daitch, WA4MZZ
Louisville, Georgia

wire sizes

Dear HR:

I noted in Forrest Gehrke's article ("A Precision Noise Bridge") in the March issue of *ham radio* that you have religiously converted English units to metric. This has become a popular custom in many publications even though such a procedure is frequently cumbersome, and sometimes even improper.

One good example is in converting wire gauges (the machine tool industry dropped that useless "Micro" thirty years ago) to a dimension in millimeters. Giving a wire size as "No. 24" is a wire size, but 0.5 mm is not a wire size. To be perfectly explicit, the size should be given as, for example, No. 24 AWG (for American Wire Gauge), or B&S for Brown & Sharpe, or Birmingham, or Stubs, or Washburn & Moen, or Imperial, to name a few).

If you want to be completely confused, take a look at some wire tables such as those in the *Handbook of Chemistry and Physics*. While you are at it, you might notice that successive wire size numbers are 1 dB (voltage) apart. Thus if you can remember that No. 16 B&S is 0.05083 inch (0.1291 cm), you can figure out the other sizes with fair accuracy.

But please, don't confuse wire gauges with the metric system. They are simply a preferred number system based on a logical progression of diameters no matter what units are used.

Donald E. Williamson, K4HVI
Miami, Florida

data bandwidths

Dear HR:

In the article "Data Bandwidths Compared" (December, 1982, *ham radio*) the suggestion that phase modulation permits transmission of data at a rate faster than the corresponding receiver bandwidth is very misleading. Granted the sidebands may be more than 15 dB down, but those sidebands become increasingly important as the data rate goes up. The author notes that the error rate is higher; I wonder whether he considers 50 percent errors satisfactory for 800 BPS transmission through a 400-Hz channel.

A more useful study would have included bit sequences other than the (1,0,0,1,0,1,1,0) in the examples. Though this may be random, it is by no means the only possible eight-bit sequence. Each has a different spectrum. A proper analysis would have found the average spectrum from *all* possible eight-bit sequences.

This article had a lot of potential. I'm afraid it will leave many readers with some incorrect ideas about data rates and channel bandwidths.

Dick Simpson, W6JTH
Palo Alto, California

the battlefield

Dear HR:

The editorial "The Battlefield" by K2RR calls for respect for the DXers' right to 3795-3800 kHz. Before many Amateurs could respect special DX frequencies, they would first have to respect the wham-bam, touch-and-go-type of contacts. Today's DXing is very disappointing for those Amateurs who have known DXing as a means of gaining a personal or meaningful acquaintance with foreign hams. Too many so-called DXers, I believe, have only graduated from matchbook collecting, and their DXing techniques do not deserve any special respect.

Warren U. Amfahr, WØWL
Des Moines, Iowa

good job

Dear HR:

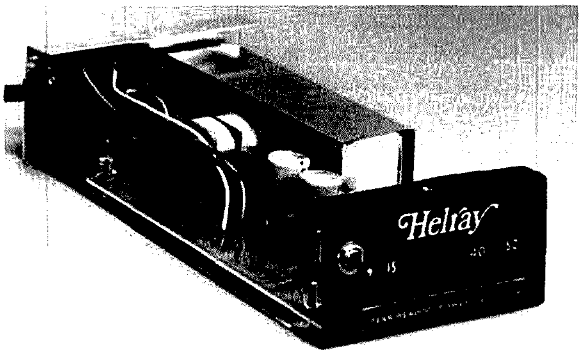
I'm writing to let you know that I'm renewing my subscription because of your February issue. I hope you keep having articles on UHF/microwave equipment. There are too many general-interest articles and magazines. Keep up the good work.

Denney Pistole
Port Hueneme, California

measurement of
PEP output power

Instantaneous indication of peak levels with a multi-lamp LED display

Ever since single sideband first became popular among Amateurs in the early 1960s, a problem has existed which has plagued both the Amateur and the regulating authorities. The problem has been to accurately measure the output power of an SSB transmitter when actually on the air — especially when operating near the maximum legal limit. Most countries specify a maximum power level that must not be exceeded by the Amateur. In the U.S.A., a limit is set on the allowable dc input power to the final amplifying devices — be they valve or solid state. In Australia, the United Kingdom, and many other countries, a limit is set on the allowable peak rf output from a transmitter.



Interior view of a 50-watt model PEP indicator. The assembly slides into an extruded aluminum case. The sampling head is contained in the rectangular housing at the rear. All other components, including the power supply, are mounted on an L-shaped PCB.

While measurement of peak output power has much to commend it in terms of the extra freedom it gives to the designer of the transmitter, the problem of accurately measuring this peak output power persists. The Australian and British authorities recommend use of an oscilloscope and a two-tone oscillator to establish the permissible limit of 400 watts PEP output, and then impose a requirement on the individual Amateur not to exceed this limit when in actual operation. This method is, at best, clumsy, and it has the additional drawback that the recommended equipment is expensive. Furthermore, the oscilloscope method really requires a long-persistence tube if any worthwhile degree of control is to be exercised. It is the purpose of this article to describe a simple device which can be used to accurately measure peak output power and which uses a dc voltage for calibration purposes.

Consider first the completely conventional method of measuring rf voltages. **Fig.** 1 is typical of the sort of "rf voltmeter" that has been described in the literature for decades. A diode in series with a capacitor is placed across the load. Provided the source gives a steady rf output, the capacitor then charges to the peak value of the rf waveform and this voltage can be measured with a normal, high-impedance dc voltmeter.

However, in the case of SSB, the output is not steady but varies at a syllabic rate. The normal mov-

By H.L. Hepburn, VK3AFQ, 4 Elizabeth Street, East Brighton, VIC, 3187, Australia

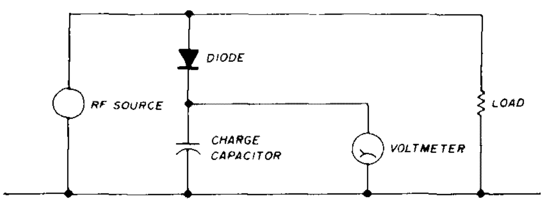


fig. 1. Conventional rf voltmeter.

ing coil or taut band meter movement has considerable mechanical inertia and is not able to follow fast variations in level. Indeed the meter will act as a sort of integrator and, to an extent which is set by its mechanical construction, will tend to average out the peaks of the speech waveform. Under these circumstances, it is just not possible for a mechanical device to register the fast peaks which cause over-limit operation. (Editors' note: The manifestations of which can be splatter, broad signals, and bad feelings on a crowded band.) This deficiency in mechanical meter movements makes the measurement of peak speech power impossible regardless of whether the license limits are defined in terms of input or output power.

What's needed is a device that gives an accurate and unambiguous indication when license limits have been exceeded. Calibration of such a device should preferably be by way of easily measured dc voltages.

Before I describe such a device, a few relationships are provided as background: The relationship between the voltage on a feedline having a 1:1 SWR and the peak output power into the load is given by the formula:

$$E = \sqrt{2PR}$$

where E is the instantaneous voltage on the feedline;
P is the peak power in watts; and
R is the load resistance in ohms.

In the case of the conventional 50-ohm antenna system, this expression simplifies to:

$$E = 10\sqrt{P}$$

Working with the Australian "allowance" (that is, maximum power limit) of 400 watts PEP output, the voltage on the 50-ohm feedline must not exceed:

$$10\sqrt{400} = 200 \text{ volts}$$

Thus, to meet the Australian license requirements, this 200 volts on the 50-ohm feedline must not be exceeded at any time.

threshold circuit

Fig. 2 is a circuit diagram of a simple over-limit indicator. It shows whether the voltage on the 50-ohm feedline has or has not, even momentarily, exceeded 200 volts (or 400 watts PEP) into a 50-ohm load. First, the voltage on the line must be reduced in a controlled fashion so that the capabilities of the diode are not exceeded. A resistive divider is used, which, at dc only, has a divide-by-5 ratio. At rf this ratio can still be maintained with the addition of several compensating components. The method of compensating the divider is detailed later.

At the output of the primary divider, the 400 watts peak on the feedline is now represented by:

$$\frac{200}{5} = 40 \text{ volts}$$

Forty volts is still a bit high for solid-state devices, so a secondary divider consisting of a 5.6 megohms in series with a 1.2-megohm fixed resistor and 1.0-megohm trimpot is used. This secondary divider has a

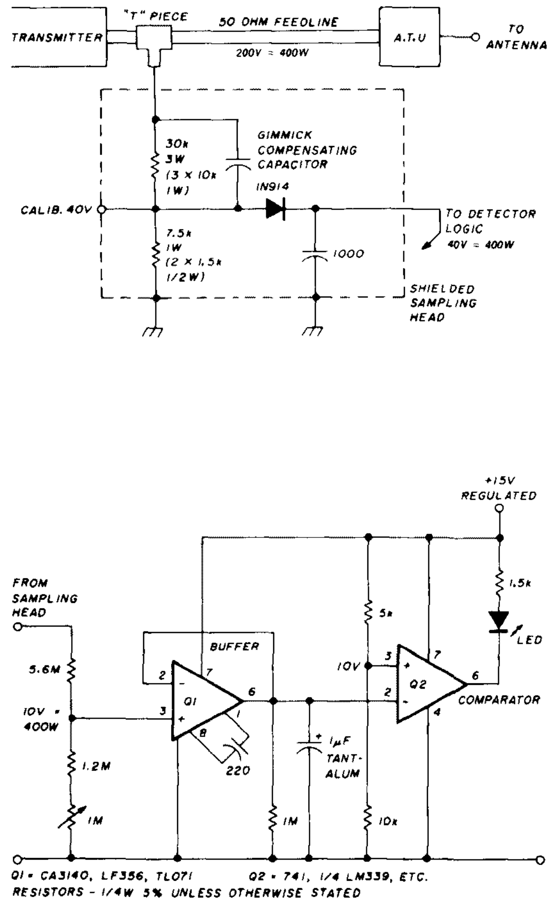


fig. 2. Single-stage indicator.

nominal ratio of 4:1 so that when it's adjusted, the original 200 volts on the feedline is now reduced to 10 volts. Since the output from the sampling head after rectification is audio with frequency components up to only 3 kHz, no compensation of the secondary divider is necessary.

Output from the secondary divider is buffered using a high-input impedance FET op-amp (CA 3140, LF 356, TL 071, etc.) connected as a voltage follower. Output from the buffer is applied to the inverting input of an op-amp set up as a comparator. The noninverting input is held at exactly 10.0 volts by means of the resistors across the 15-volt regulated supply. With no voltage input from the feedline there is no voltage on the inverting input of the comparator. The output of the comparator will be close to the 15-volt supply line and the LED does not draw current.

As soon as the 50-ohm feedline exceeds 200 volts (equal to 400 watts PEP), 10 volts or more is applied to the inverting input of the comparator. The comparator output drops, and the LED lights indicate that the allowable limit has been exceeded. For very short voice peaks, the eye may not register the fact that the LED was on. A rudimentary pulse stretcher consisting of a 1-megohm resistor and 1.0-μF tantalum capacitor is attached to the output of the buffer to ensure that there is a minimum on time for the LED. Calibration of the device requires a dc voltage source variable around 40 volts.

The sampling head is removed from the feedline and the variable dc supply applied to the **CALIB 40 volts** point on the primary divider. An accurate voltmeter is used to adjust the calibrating supply to exactly 40.0 volts. The 1.0-megohm trimpot is then adjusted so that the LED just lights. The variables in the system (voltage drop across the diode, offset voltage of the op-amps, inaccuracies in the resistors, and so forth) are all nicely taken care of by this dc calibrating procedure.

Physically, the sampling head and primary divider may be separate from the comparator logic and should be fully shielded. A 5 × 1¼ × 1¼ inch (135 × 32 × 32 mm) box made of double-sided circuit board makes a very simple shielded enclosure for the dropping resistors, the diode, the charge capacitor, and the calibrating voltage input point. The rf input can be a standard SO-239 socket, the rectified output an RCA socket, and the calibrating voltage input a standard banana plug socket. **Fig. 3** shows how the head is assembled. A suitable power supply incorporating both logic and calibrating voltages is shown in **fig. 4**.

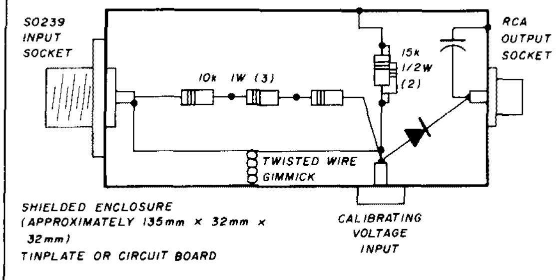


fig. 3. General arrangement of sampling head.

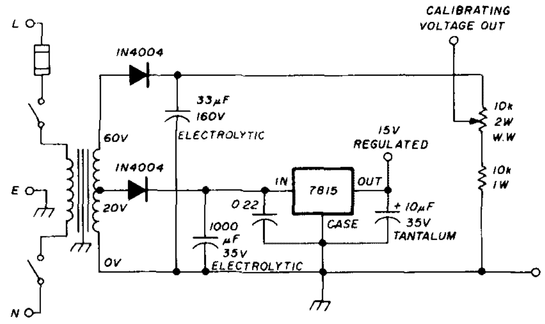


fig. 4. Power supply schematic.

sampling head compensation

The rectifying diode in the sampling head has a small but finite series capacitance. This capacitance is effectively across the 7.5-kilohm base resistor of the primary divider. At all significant rf frequencies this capacitance has a reactance which effectively reduces the 7.5 kilohms to a considerably lower value depending on frequency. Unless something is done, the 5:1 ratio of the primary divider no longer holds.

If a compensating capacitor is placed across the three 10-kilohm resistors forming the upper half of the primary divider, and adjusted so that its capacitance is exactly one quarter of the diode capacity, then the division ratio will remain at 5:1 although the effective resistance values may be significantly different from the dc resistance values. The compensating capacitor is a "gimmick" formed by attaching two insulated wires, one to the input socket and the other to the diode/7.5 kilohm junction, and twisting them together as shown in **fig. 3**. The test set-up for adjusting the compensating capacitor is given in **fig. 5**.

calibrating the sampling head

The station transmitter in the CW mode is operated into a standard rf wattmeter. This wattmeter need not be a high-power one — 10 to 20 watts is adequate. The sampling head is T'd into the line connecting the transmitter to the wattmeter. The transmitter is set to 3.5 MHz and the carrier level control

table 1. Eight-level indicator voltages and resistor values.

A	B	C	D	E	F	G	H	I
power level PEP watts	peak voltage on 50-ohm line	voltage out of first divider	voltage out of second divider	fixed dc voltage required on noninvert inputs	required ladder resistor value ohms	make-up of ladder resistors by paralleling	actual value of ladder resistors ohms	percentage error
					R9-3820	39, 180K	3817	−0.08
500	223.61	44.721	11.180	Q8-11.180	R8- 573	620, 8.2K	576	−0.52
450	212.13	42.426	10.607	Q7-10.607	R7- 607	620, 27K	606	−0.16
400	200.00	40.000	10.000	Q6-10.000	R6- 646	680, 12K	643	−0.46
350	187.08	37.416	9.354	Q5- 9.354	R5- 694	820, 4.7K	698	+0.58
300	173.21	34.641	8.660	Q4- 8.660	R4- 754	820, 9.1K	752	−0.26
250	158.13	31.623	7.906	Q3- 7.906	R3- 835	910, 10K	834	−0.12
200	141.42	28.284	7.071	Q2- 7.071	R2- 947	1.0K, 18K	947	—
150	122.47	24.495	6.124	Q1- 6.124	R1-6124	6.2K	6200	+1.2

advanced until a suitable reading — say 10 watts — is registered on the wattmeter. The absolute accuracy of the wattmeter is unimportant. It is only necessary that the transmitter output be set to a fixed value during the calibration procedure. Now read the value on the voltmeter attached to the output socket of the sampling head.

The transmitter is switched to 29.5 MHz and the carrier level advanced until the same 10-watt output is registered on the wattmeter. The voltmeter is again read.

If the 3.5-MHz voltmeter reading is greater than the 29.5-MHz reading, there is insufficient capacitance in the "gimmick" and more twists should be added. If the 29.5-MHz reading is higher than the 3.5-MHz reading, the head has been over-compensated for, and the "gimmick" should be untwisted. The aim is to get the voltmeter readings at both 3.5 and 29.5 MHz to be the same. Note that putting the cover on the sampling head enclosure has a slight effect on the readings and should be allowed for.

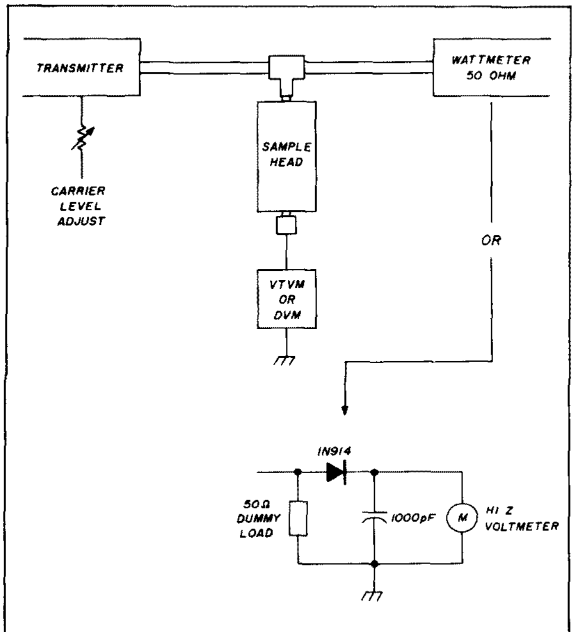


fig. 5. Test setup — head compensation.

additional level indicators

The single LED indicator described above serves one fundamental purpose, to show whether a set power level on the feedline has been exceeded. Other levels can be set simply by changing the dc voltage on the comparator noninverting input.

A useful and practical extension is to use a multiplicity of comparators, whose inverting inputs are fed in parallel by the buffer and whose noninverting inputs are set at different dc levels, each dc level corresponding to a specific rf output power level. **Fig. 6** gives the schematic of such an arrangement suitable for eight levels. The resistive ladder (R1 to R9) is calculated on the basis of 1000 ohms per volt. In most cases it will be necessary to parallel two standard 5-percent resistor values to obtain the (usually odd!) values required by each "rung" of the ladder. The appendix details the method of calculating resistor values.

However, the internal resistive ladders are characterized for either a linear or logarithmic relationship between input and comparator trigger points and are not applicable when the input/output voltages have a *square root* relationship, as is the case when power is being measured. By using devices such as the National LM339 (which have four comparators in a single 14 pin DIP), and an external resistive ladder the

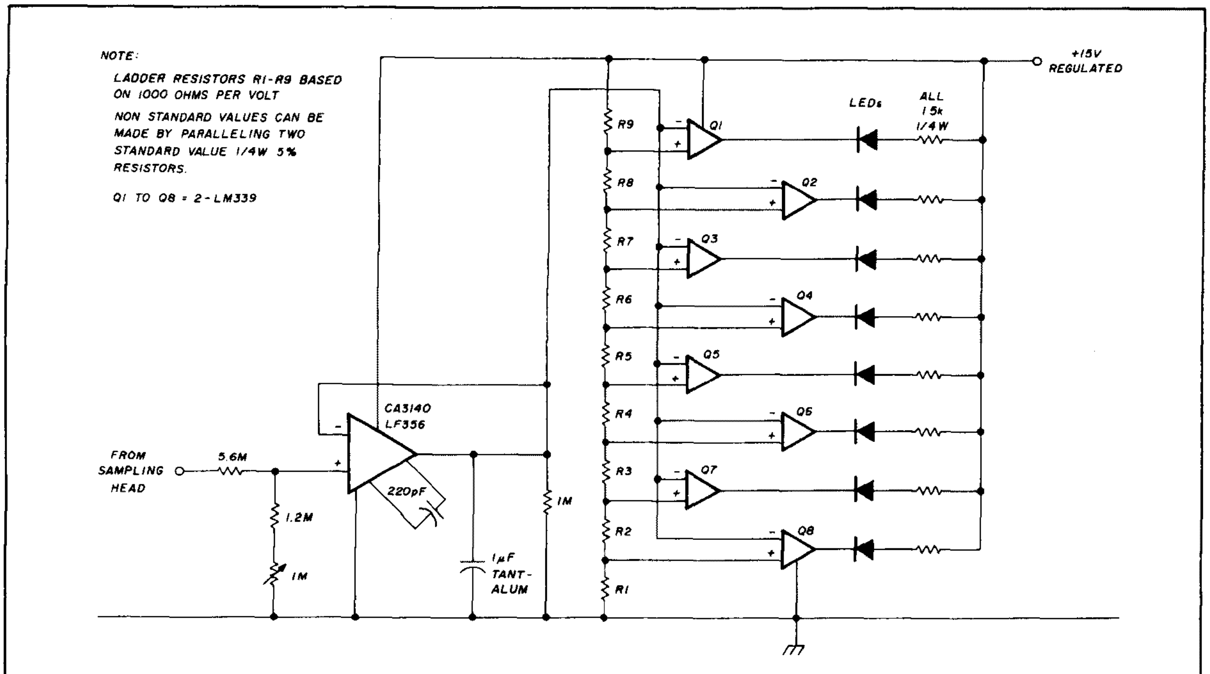


fig. 6. Eight-level indicator.

design can be made very flexible and varied to suit most applications.

In practice, an eight-step indicator covering 150-500 watts in 50-watt steps is a very useful device. Using green LEDs up to the 350-watt mark and red LEDs for the 400, 450 and 500 watt levels, the green/red transition is very obvious. It is necessary only to place the indicator lights where they will be easily seen. It's not necessary to watch the indicator; you will notice the change to the red LED even if you are not consciously watching for it. The device was developed as a means of showing when the allowable output power is exceeded during actual "on air" operation. For this purpose the stepwise approach is more than adequate.

For tune up purposes, using a carrier only, there is no doubt that a continuous indication is advantageous. The simple addition of a high impedance voltmeter (10,000 ohms per volt or better) between the output of the buffer op-amp and ground accomplishes just this, since, at that point, the voltage varies exactly in step with the voltage on the feedline. The voltmeter 0-15 volt range should be calibrated according to the power output being used. Fig. 7 shows how this is done.

conclusion

A commercial version of this indicator, based on the principles outlined above is presently being marketed in Australia. More importantly, the Australian regulatory authorities have tested the device and have pronounced it as an acceptable alternative to the official oscilloscope method. The three standard ranges used in VK are 5-40 watts PEP in 5-watt steps, 25-200 PEP in 25-watt steps, and 150-400 PEP in 50-watt steps. Additionally, many "specials" have been made for other than Amateur use with 2 kW PEP being the maximum to date. The only changes necessary to do this have been in the values of the primary divider resistors and their wattage. The use of 5-percent-tolerance resistors in the unit has proved to be entirely satisfactory.

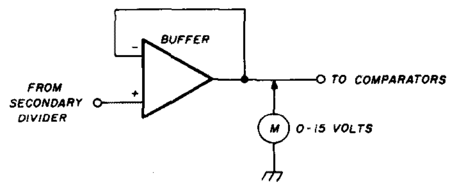


fig. 7. Addition of tune-up metering.

appendix

Calculation of ladder resistor values: **table 1** includes data for an eight-step indicator covering 150-500 watts PEP in 50-watt steps. It can be revised for other ranges and other spacings by the following procedure. A supply of 15.0 volts regulated is required.

1. List in column **A** the power levels and spacings required

2. List in column **B** the peak voltage on a 50-ohm line for each power level calculated from the formula $E = 10\sqrt{P}$

3. List in column **C** the voltages expected out of the 5:1 primary divider for each power level calculated from the formula $E_{prim} = 0.2E$

4. List in column **D** the voltages expected out of the 4:1 secondary divider for each power level calculated from this formula:

$$E_{sec} = 0.25\,E_{prim}$$
$$\text{or } E_{sec} = 0.05\,E$$

Note that the E_{sec} voltages calculated in step 4 are the trigger voltages, which have to be set on the noninvert inputs of the comparators. Call these $E_{trig}\,Q1$ to $E_{trig}\,Q8$ and enter in column **E**.

5. The resistance required between ground and the noninvert input of $Q1$ is given by the formula $E_{trig}\,Q1 \times 1000$. This resistor is shown in **fig. 6** as $R1$. If the resistance value is not within 2 percent of a standard resistor value, two resistors will have to be paralleled to get the correct value.

(a) Let R_a be the value required

(b) Choose the next standard value of resistor higher than R_a. Call this Resistor R_s

(c) Calculate the value of the paralleling resistor required with this formula:

$$R_{par} = \frac{R_s \times R_a}{R_s - R_a}$$

This calculation will usually give a result which is not a standard value but which is between standard values. Again, choose a standard value for R_{par} which is the next one above the calculated value of R_{par}.

(d) Check that R_s paralleled with R_{par} is within 1 to 2 percent of the required value of R_a using this formula:

$$R = \frac{R_s \times R_{par}}{R_s + R_{par}}$$

Enter the resistor values to be used and their effective parallel value in columns **G** and **H**.

6. The value of R_2 is given by the formula

$$(E_{trig}\,Q2 - E_{trig}\,Q1) \times 1000$$

To realize the correct resistance value, steps 5(a) and 5(d) are repeated.

7. The values of $R3$ to $R8$ are calculated much like R_2, using the general formula:

$$(E_{trig}\,Qx - E_{trig}\,Qx - 1) \times 1000$$

and steps 5(a) to 5(d) as before.

8. The value of $R9$ is calculated from this formula:

$$(15.0 - E_{trig}\,Q8) \times 1000$$

Note that if the actual voltage available on the regulated 15-volt line is significantly different from 15.0 volts (say, over 15.3 volts or under 14.7 volts) then calculate $R9$ using this formula:

$$(Actual\ volts - E_{trig}\,Q8) \times 1000$$

(a) Calculate the percentage variation of the used values of R1 to R9 as shown in column **H** from the theoretical values of R shown in column F. Provided they are within 2 percent, no change is required. As **table 1** shows, the error in each step is likely to be very much less than 2 percent and of no practical significance.

ham radio

design of short vertical antennas for the low bands: part 2

An efficient radiator for 160 and 75 that uses two top hats

a good radial ground system in a small back yard

You don't need a large back yard to have a good radial ground system. I helped a friend install a good system in a back yard that measured only 65 × 40 feet (20 × 12 meters). At that location there are chain-link fences around the entire yard, and about thirty fence posts. Other fences terminate near the corners, but they are not bonded to the back yard fence. We placed an aluminum plate on the grass, approximately in the center of the yard, and each fence post was connected to that plate by a No. 17 galvanized steel wire. The wires were buried in a slot dug by an ordinary edger. That made thirty radials, the longest of which was not more than about 40 feet (12 meters).

The fence posts and the plate were drilled, tapped, and provided with solder lugs, and the wires were soldered at each end. A temporary base-loaded whip was then resonated to 1.850 MHz, and the impedance was measured with a noise bridge. During these measurements, various fence corners were temporarily connected together with clip leads, and the effects were immediately noticeable. Within an hour, the ground system measured a fairly reliable 10 ohms. No measurement was performed at 3.8 MHz; however, it certainly would measure less than 10 ohms. Note that this minimal, small-lot radial system is *better* than the hypothetical ground used as the basis for calculations in this article.

The first half of this article reviewed the characteristics of several short verticals over ordinary ground radial systems. This information is now applied to the *construction* of a very short two-band trapped vertical.

the antenna system

The antenna *consists of two top hats separated* by a parallel resonant 75-meter trap and a 35-foot-long (10.67-meter-long) section of 2-inch aluminum irrigation pipe. It resonates on 1.836 MHz and 3.830 MHz. The 75-meter top hat comprises four 8-foot (2.44-meter) pieces of 3/4-inch aluminum tubing, mounted at right angles to themselves and to the mast.

Four 50-foot-long (15.24-meter-long) sections of No. 8 aluminum clothesline wire double as the top guys and the 160 meter top-hat. The trap is placed between these wires and the top of the mast. For design purposes the trap inductance is 4.5 μH with an equivalent inductance when paralleled with 400 pF of slightly over 5 μH at 1.8 MHz. With the trap shorted the antenna resonates at 3 MHz.

The general layout is as shown in **fig. 1**. Six rather small trees and the lower guys are not shown. **Fig. 2** shows the electrical connections.

Since 2-inch irrigation pipe is very limber, it needs to be guyed. The first installation had eight guys; four comprised the 160-meter top-hat, plus four more at the 60-percent level. On one windy day the mast section below the lower guys vibrated at 20 to 30 Hz, displacing at least half an inch. Wind velocity at the time was probably 30 to 40 mph (50-65 kph). Another set of guys was subsequently installed at the 12.5-foot (3.8-meter) level. The positions are now at 35 feet (10.7 meters), 21 feet (6.4 meters), and 12.5 feet (3.8 meters). This damped the motion.

By W.J. Byron, W7DHD, 5 Lambert Lane, Robbinsville, New Jersey 08691

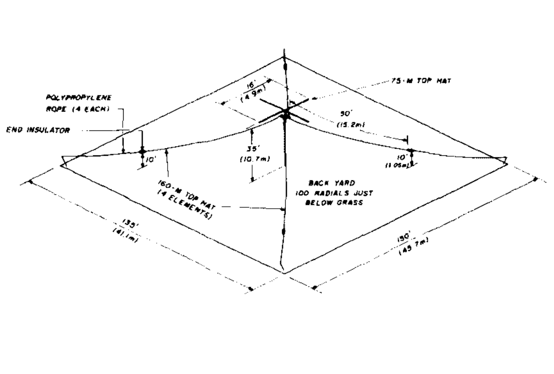


fig. 1. Configuration of trapped 75/160 meter vertical, approximately to scale.

The system is resilient enough so that continued vibration certainly would not have damaged the structure, but it does stranger things: it unscrews nuts, bolts, and machine screws, and it causes fatigue in electrical connections. The fastening points for the guys on the mast are steel devices designed for connecting 2-inch fence posts together. All guys are broken up into sections approximately 10 feet (3 meters) long, insulated by TV-mast-type "egg" strain insulators.

A 30-kV ceramic stand-off insulator is used, supporting the weight of the antenna system. In this application the potential across it is under 100 volts. For this (and even more stringent requirements) an empty (but corked) champagne bottle will do just as well: they have heavier walls than soft-drink or beer bottles. An alternative base insulator for this antenna could be a 500-ohm "glo-bar" resistor, since the higher antenna input resistance (achieved on 75 meters) is only 19.6 ohms.

It is a much different story, however, if a *base loading* coil is either wound around the insulator or connected across it. Assuming total base-loading, the rf potential on a $\lambda/16$ vertical could reach 5000 volts, and that puts insulator requirements in an entirely different light.

the ground system

Short verticals must have good rf grounds. At my location the longest radial wires are the four diagonals that run across the back yard, each about 90 feet (28 meters) long. The shortest are the two that span the short side, each 65 feet (20 meters) long. All other radials are of lengths between those limits. Since all are shorter than $\lambda/4$, it was decided to put as many radials down as would be practicable. One hundred radials were installed, utilizing 7,300 feet (2250 meters) of galvanized steel, 17-gauge electric fence wire.

Heeding the advice of Jerry Sevick, they were laid down, in October, 1978, on the surface of freshly close-cut lawn. At those places where the ground dipped a wooden stake was driven in flush with the average ground height, and the radial wire was stapled to the top of the stake. None of the radials were buried. By midsummer, 1979, grass had concealed all the wires. By that time a metal detector would have been necessary to find even one of them. A couple of wires did catch in the mower in the spring of 1979, but these two were replaced and there has been no trouble since.

The radials are attached to the base plate by solder lugs and No. 6-32 (M3.5) stainless steel screws. The base plate is an 18-inch (0.5-meter) square piece of 6061-T6 aluminum, 1/4 inch thick. One hundred holes were drilled with a 6-32 (M3.5) tap-drill (twenty-five on a side) and then tapped by hand.

Most radial wires are terminated by an 18-inch (0.5-meter) piece of reinforcing rod, supplied and cut by a local building supply company. Some are terminated by 4-foot (1.2-meter) ground rods. These were reserved for the shortest radials. All this paid off; the ground loss was very small.

lightning protection

Something should be said about grounds for lightning protection. At my location some of the larger trees and also some of the power-line poles are higher than the antenna, but none of them are in the immediate vicinity of the antenna. I drove four 4-foot (1.2-meter) rods below ground level several feet off the corners of the baseplate and connected them to the corners by a No. 8 solid copper wire. The rods are on about 6-foot (1.8-meter) centers.

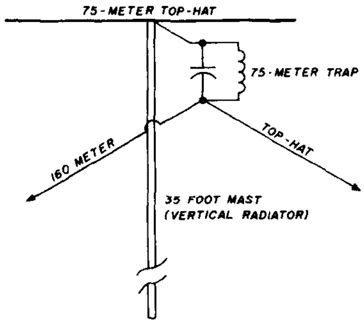


fig. 2. Top-hat electrical connections.

A little research into grounding revealed that if two rods are spaced close together, say less than a foot or two, their parallel resistance to ground is no less than that of one rod. In other words, rod surface area alone has very little to do with contact resistance. Ground rods should be spaced at a distance *at least* equal to their length for minimum resistance to ground. It has been shown also that athough there is little correlation between their surface areas (rod diameter) and resistance, there is a *direct* correlation with their length. So the proper lightning ground should have been composed of 8-foot (2.5-meter) rods spaced approximately 10 feet (3 meters) apart.

Even if the ground resistance could be reduced to 5 ohms (ohmic resistance, dc), a direct hit delivering a pulse current as low as 2,000 amps would still produce a voltage at the base of the antenna of 10 kilovolts with respect to nearby objects such as coaxial feedline and the shack. Lightning ground rods can help protect the shack or the house; *they will not protect the equipment.* Despite the rods, when I leave the shack for the day I uncouple the coaxial transmission line and leave it dangling on the wall.

matching

The system exhibits two distinct resistive feedpoint impedances, both less than 50 ohms. Several options exist for feeding the antenna, but the one I chose involves running a "flat" transmission line (that is, matched at both ends), the impedance transformation being accomplished by a transmission line transformer (TLT) at the base of the antenna. Construction details are given in **fig.** 3. There is a slight mismatch at the 75-meter tap. The load impedance (measured) is 19.6 ohms. At the 2/3 point of the TLT, the impedance is 22.22 ohms, providing a best-case match (VSWR) of 1.13. The 160-meter tap is exactly at the measured value of antenna resistance, 5.5 ohms, which occurred by lucky coincidence. Those two taps could be switched by a relay energized in the shack. At the time I didn't have a relay with sufficient contact force or size to do the job. At the kilowatt level, the base current in the 160-meter antenna is nearly 13.5 amperes. At this level each 0.1-ohm circuit resistance dissipates 18 watts, a loss which would hardly be missed in the resultant signal but which is serious if it happens to occur in a relay contact, which may have a total volume of less than one-tenth of a cubic centimeter.

Another option is to use a quarter-wave matching section, computed in the usual way:

$$Z_M = \sqrt{Z_i Z_o}$$

For $Z_i = 50$ *ohms*, and $Z_o = 5.5$ *ohms*, $Z_M = 16.6$

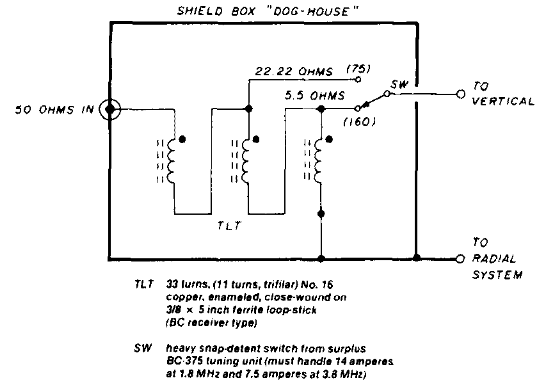


fig. 3. Transmission-line transformer (TLT) and switch.

ohms, almost exactly one-third of 50 ohms. It would require a total of 350 feet (107 meters) of RG-58: three cables, each 117-feet (35.6-meters) long and operated in parallel. That is a lot of cable but it works. On 75 meters, the length would approximate $\lambda/2$, and the 19.6-ohm resistance would be repeated at the transmitter with an intermediate point SWR on the 16.6-ohm transmission line as low as 1.18. Its use would require a transmatch at the transmitter, however.

When it's cold outside it's a chore to go out and switch the TLT. I soon discovered that with the TLT on the 5.5-ohm tap and the excitation on 3.8 MHz, the resultant 4:1 SWR could be compensated for by the transmatch at the transmitter with apparently no loss in efficiency. The loss due to that mismatch is about 1 dB. Whenever the weather is foul enough to discourage a "switching trek" I use the transmatch. Note that the loss is low only for frequencies of 4 MHz and lower. It would not be true at 30 MHz.

Figs. 4 and 5 are plots of the measured VSWR within the two bands. The measurements were taken from the shack at the end of the long RG-8/U transmission line. I used a Bird Model 43 wattmeter and the barefoot exciter with its digital frequency display. The readings reveal a couple of important facts:

1. The measured impedances at the antenna base were correct.

2. The transmission line transformer is adequately designed.

The antenna systems, although narrow in bandwidth, can be tuned by means of transmatches. I use two, one for 160 meters and the other for 80 through 10 meters, enabling me to operate over a range of 200 kHz on 75 meters and 100 kHz on 160 meters

with little loss. Mine are both the McCoy "ultimate transmatch" type. In general, transmatches are recommended for use with trapped systems to reduce harmonic radiation.

performance

The 75/160 meter antenna has proven to be an effective radiator. Because of the low-angle radiation and the efficiency (high radiated field), I can make two-way contacts on 160 meters at high noon in mid summer over distances of up to 200 miles. At the equinoxes, I've made two-way SSB QSOs with New Zealand. At various other times in the winter the system has enabled me to work England, Brazil, Panama, Nova Scotia, Bermuda, and most of the United States. On 75 meters the performance is much the same; I've made contacts with all of the above places plus Germany, Alaska, and western Australia. All of these contacts were incidental, since I do not seek DX actively.

conclusion

In various contacts I've made with other users of verticals, it's become obvious to me that the Amateurs who have become disillusioned with, and consequently abandoned, short verticals did not fully understand just how important the ground system is. The shorter the antenna, the more sensitive it is to ground losses. The problem becomes worse when the antenna is both short and base loaded. Consequently, more attention should be paid to the ground system than to the "top-works," while not neglecting the reduction of ohmic losses in coils, relay contacts, electrical connections, and bonding strips.

I also often hear it said that the antenna should be erected over a high water table: "The water table is only five feet below the surface here, and I have

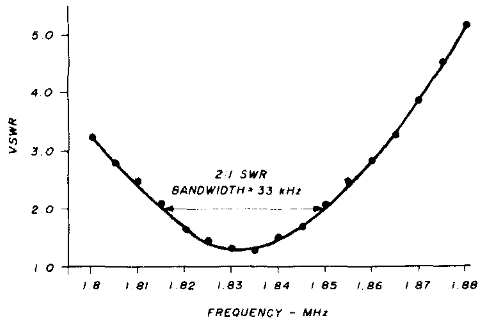


fig. 4. VSWR plot between 1.800 and 1.880 MHz of the 160-meter system.

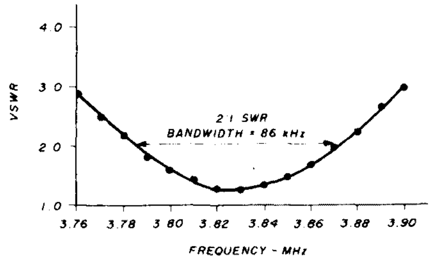


fig. 5. VSWR plot between 3.76 and 3.90 MHz of the 75-meter system.

grounding rods driven right down into it." That means absolutely nothing, unless that "ground water" is salt water. Fresh water, particularly if it is potable, has the conductivity of *poor earth*. The answer to all this, of course, is to know what the antenna should "look" like, and to measure the base-impedance over the ground system. If the system should present a 20-ohm load and has no ohmic loss but measures 50 ohms, you know immediately that the ground resistance is 30 ohms. Consequently, the signal from that antenna will be considerably reduced, *even though it will present a very good match to the transmission line*.

The use of a very short base-loaded whip (such as a mobile whip) reveals something interesting. It is such an inefficient radiator that it can be used to estimate ground loss. Resonate the whip over the ground system and measure the total resistance. Then remove the whip above the coil and replace it with a variable capacitor to ground. Reresonate the coil.

When all this is done, the radial system is effectively removed from the circuit. You can then make another noise-resistance measurement, determining the coil resistance. Absence of the whip ($R_r \cong 0$) has no measurable effect. The ground resistance is then the difference between the two measurements.

acknowledgements

I wish to acknowledge the encouragement I received from W1DB, the late Nick Lefor, who persuaded me to write this article. I also wish to thank Tony Sivo, W2FJ, Walter Schulz, K3OQF, and Nevell Greenough, N2GX, for reading drafts and for other assistance; also Jerry Sevick, W2FMI, from whom I learned to make my first transmission-line transformer; and Edmund Laport, my principal reviewer.

ham radio

vertical phased arrays: part 2

Part two examines array siting, field plot calculations, and minor lobe determinations

The theoretical design of vertical phased arrays will be the subject of this article, the second in this series on vertical antennas. Every designer must balance his performance requirements for high gain and F/B (front-to-back ratio) against his resources (space, money, and time). There is always a strong temptation to skip past the theoretical work and proceed with the more engrossing task of construction. But after having relocated the elements of an array (complete with one-hundred radial ground systems) more than once, I can tell you that a few thoughtful hours spent on design can save many hours of wasted construction time.

For example, you might want to start with a two-element array. But before you clear a site of brush and trees, consider what you might be faced with should you later decide to add elements to this array. And consider the directions of the main lobe of the changed array. For instance, a two-element array has a main lobe whose half-power beamwidth is 180 degrees. Before you decide to aim this array toward Europe or Japan remember that the signal loss sustained in orienting this array exactly east and west is about 1/2 dB (down from Europe or Japan) and only 3 dB — half an S-unit — down in a north or south direction (see **figs. 3** and **4**)*. I live in a wooded area and, not being willing to become involved in a lum-

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*Charts as introduced in the text are not in sequence. However, they are grouped as follows: two-element arrays, **figs. 3** and **4**; three-element arrays, **figs. 5** through **11**; four-element arrays, **figs. 12** through **17**. **Editor.**

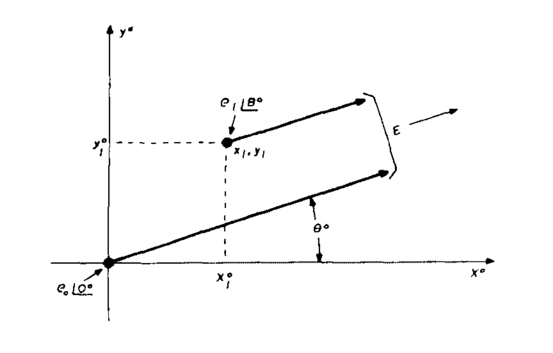

fig. 1. Generation of cumulative field in the θ direction by the two sources $e_0 \angle 0$ and $e_1 \angle B$.

bering operation, I had to strategically locate a 4-square array among the trees. Because I failed to give some advance thought to these considerations, relocations of elements were required with every addition to my array.

In a previous article,[1] I discussed some of the reasons for less-than-anticipated performance in vertical phased arrays, particularly in front-to-back *ratio*. The major fault in many designs is a failure to consider the real and significant effects of *mutual* coupling between elements; neglecting these terms can result in an incorrectly designed feed network.

symmetry is essential

Knowing the characteristics of the array, we can ensure that the correct *current* magnitude and phase exist at the input terminals of each element — for a particular direction.

However, in a switchable array the magnitude and phase drive current requirements to a specific element will depend on position (direction) chosen. Conversely each element must perform correctly for each switched direction of the array. If we expect to have identical field patterns in each direction it is necessary to ensure electrical symmetry. For example, referring to the driving-point element impedances of a typical 4-square array, spaced 1/4 wavelength, with equal amplitude current ratios, phased −90 and −180 degrees we have:

$$Z_1 = 7.9 - j7.8$$
$$Z_2 = Z_3 = 35.7 - j12.7$$
$$Z_4 = 59.2 + j42.6$$

The zero degree phase reference elements driving point impedance is Z_1. The diagonally opposite element is phased −180 degrees and its driving point impedance is Z_4. The two middle elements, each phased −90 degrees present the same impedances Z_2 and Z_3. When this array is switched through its four directions, each element, in turn, assumes each of the four electrical positions in the array. For identical patterns in each direction each element must exhibit the particular driving point impedances appropriate to these electrical positions. For example, as each element is switched to occupy electrical position 4, the same driving point impedance Z_4 must be presented to the feed network. Similarly, as each element is switched into electrical position 1 it must present the much different impedance Z_1. So, instead of physically rotating this antenna, keeping each element fixed in its electrical relationship as with Yagis, we *rotate* the *electrical* relationship of the elements and keep the *physical* relationship fixed.

This is an important difference from the design of a-m broadcast arrays. Broadcast arrays are seldom switched, being designed for a particular listening area, and with departures from symmetry often intentional. For a switched array, each element's self-impedance, and each of its mutual impedances, must be as similar as possible.

Electrical symmetry is a function of the physical symmetry of the array which includes groundplanes and other nearby conductive structures. Metal towers, other antennas, guy wires, roof gutters, and leaders — that is, any conductive line within a wavelength of any part of the array, especially if it is at or near resonance (a multiple of a half-wavelength and ungrounded, or a quarter-wavelength and grounded) — should be avoided. Otherwise, a means must be found to prevent resonance. An example of preventing resonance would be to break up guy wires with insulators, making them ungrounded quarter-wave sections. When siting an array, then, look carefully around the area before starting work for anything that can act as another antenna. Unlike Yagis, where making spacing adjustments involves loosening a few clamps, low-band vertical arrays, with their groundplanes, are not easy to make adjustments on.

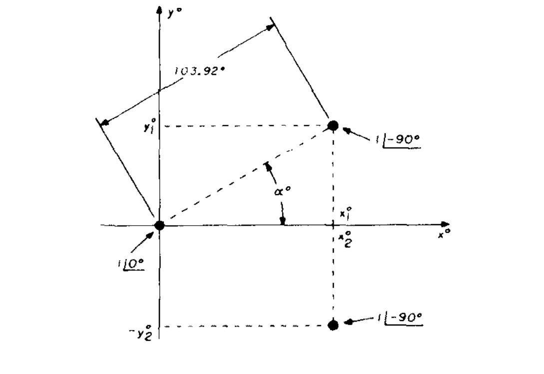

fig. 2. Example showing how the cumulative field from an equilateral triangular array can be calculated.

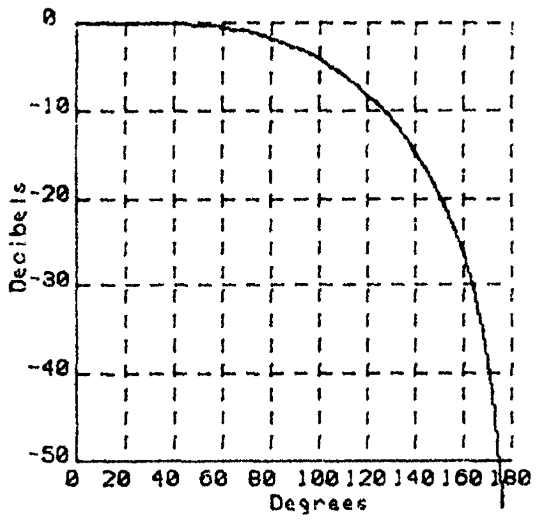

fig. 3. Two-element, 1/4-wave spacing −90 degree phasing, equal amplitude current array. [In figs. 3-17, 0 through 180° shown. *Editor*]

extensive ground systems

Don't scrimp on the groundplane. At least sixty radials a quarter-wavelength or longer should be under each element. If in some directions this is not possible, use radials at least an eighth wavelength long in even larger quantity. At those azimuths your array will probably have a higher elevation angle. However, more radials, even if short, help keep the angle down.

Theoretically, an infinitely conducting groundplane is required, but it's not practical to copper plate the neighborhood. Don't make the mistake of thinking that twenty or thirty radials is approaching the point of overdoing it! Incidentally, if you can, lay radials on the surface. If you must bury them, keep them as close to the surface as possible. Large-size wire is not necessary; I use No. 24 PVC hookup wire. Galvanized steel fence wire is not a good idea: it corrodes very quickly, becoming totally ineffective as a radial.

characterizing the array

After choosing the site it's necessary to see what kind of an array can be fitted within that area, what its characteristics might be and its switchable directions. The calculation of horizontal field patterns is an exercise in trigonometry.

Since F/B ratio is a major interest, we need to explore in greater detail the pattern in the rear area of the array. Inspection of the field equation shows that subtractive operations take place here, often resulting in small fields which have large variations with little changes in azimuth. This necessitates many azimuth calculations to reduce the granularity of the plot. Compare the similar arrays of **fig. 13**, a plot in 2-degree increments, with **fig. 16**, plotted in 10-degree increments. We would not want to miss seeing the actual variations, since deep nulls may be used later in checking out the array. A programmable calculator or a small computer is an obvious choice for handling this drudgery. When the HP-35 scientific calculator was introduced 10 years ago I plotted a three-element in-line array using 10-degree increments. The process required about 8 hours and seemed lightning-fast, but everything is relative; now I watch a Sharp PC-1500 do this in 2-degree increments in 5 minutes — including drawing a graphical representation.

multi-element array equation

The equation for the total field from any multiple element antenna is:

$$* \; E = e_0 \underline{/B_0^\circ + X_0 \cos\theta^\circ + Y_0 \sin\theta^\circ}$$
$$+ \; \ldots \; + e_n \underline{/B_n^\circ + X_n \cos\theta^\circ + Y_n \sin\theta^\circ}$$

*Choosing a common unit of dimension (e.g. degrees) for the vector terms allows simplification of the programming task.

where E is the total field term
$e_0, \ldots e_n$ are the individual term amplitudes
$B_0, \ldots B_n$ are the driving-point phase displacements with respect to the reference term

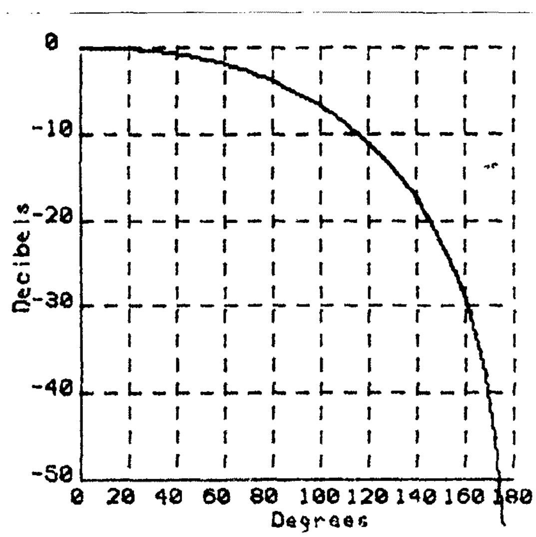

fig. 4. Two-element, 1/8-wave spacing −135 degree phasing equal amplitude current array.

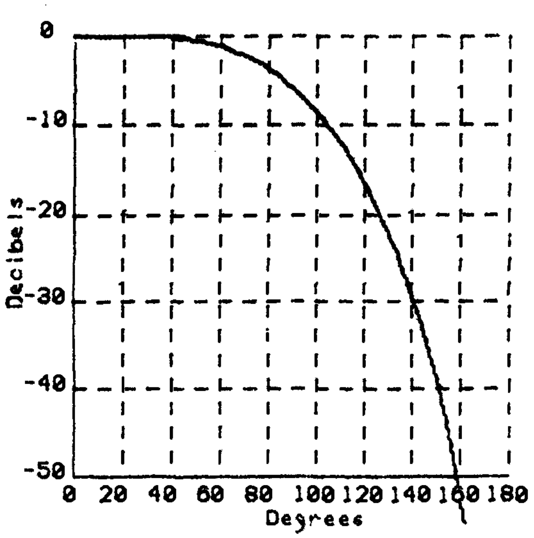

fig. 5. Three-element, in-line configuration, 1/4-wave spacing, −90 degree and −180 degree phasing, 1:2:1 current ratios.

$X_0, \ldots X_n$ and $Y_0, \ldots Y_n$ are the physical distances in terms of degrees of wavelength from the $0,0$ coordinates

θ = horizontal direction considered

There will be as many terms as there are elements. It is usually convenient to place one of the elements at the X-Y axis (origin) and to consider this element the reference element. This also simplifies calculations since all of its angular components then equate to zero. Since we are interested only in the magnitude of the vector sum of the individual terms, the angle resulting from this calculation is discarded.

two-element array calculations

Referring to **fig. 1**, consider a two-element array, with the reference element located at the origin. e_0 is the amplitude of the electric field of this element at some given distance in any direction, with a drive phase displacement B_0 of 0 degrees. Similarly, at the same distance, the field of the other element is e_1 with its driving-point phase displacement of B_1 degrees with respect to the reference element. At the given distance (assumed to be far enough removed from the array so that the combined field can be considered a plane wave) E is the vector sum of the fields e_0 and e_1 in the horizontal direction θ degrees. Note that both displacements, the physical and the electrical terms, are given in degrees.

We are interested only in determining a relative field plot for an array. We want to know what the fields are at various azimuths relative to the field at some fixed angle (usually chosen as the maximum field direction). Provided all the elements are identical, we can substitute current for voltage and we can state this current as a ratio of the reference-element current amplitude. For example, if each element were to be fed with equal current amplitude, the ratio would be 1 for each element.

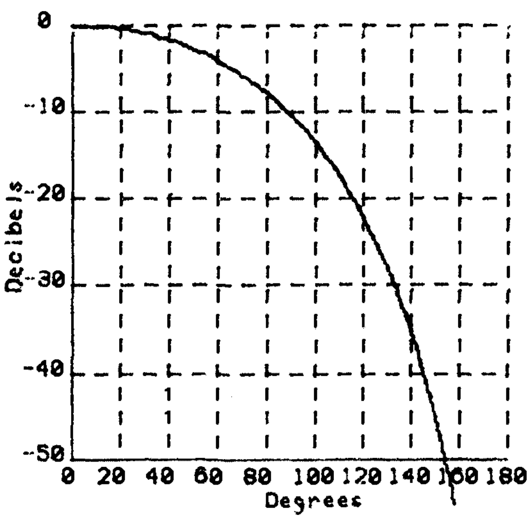

fig. 6. Three-element, in-line configuration, 1/8-wave spacing −135 degree and −270 degree phasing, 1:2:1 current ratios.

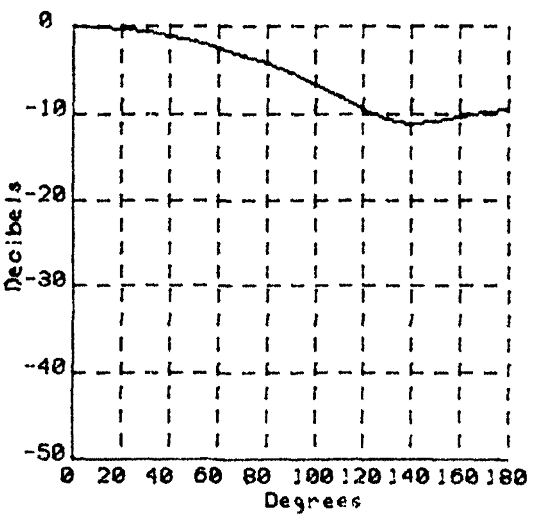

fig. 7. Three-element, triangular configuration, 0.289 wave spacing −90 degree and −90 degree phasing, 1:1:1 current ratios.

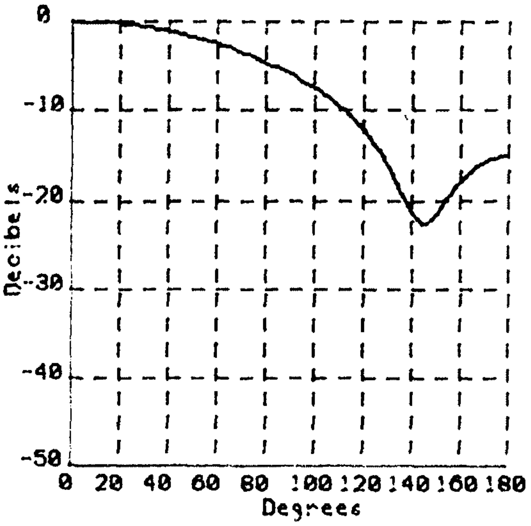

fig. 8. Three-element, triangular configuration 0.289 wave spacing −110 degree and −110 degree phasing, 1:0.5:0.5 current ratios.

three-element array calculations

A specific example illustrates how to use this equation. Referring to **fig. 2**, assume an equilateral triangular array with 0.289 wavelength spacing (that is, 103.92 degrees):

Since this is an equilateral triangle, $\alpha = 30\ degrees$.

$$X_1 = X_2 = 103.923 \cos 30° = 90°$$

$$Y_1 = 90 \tan 30° = 51.962°$$

$$Y_2 = 90 \tan (-30°) = -51.962°$$

If equal amplitude current drive feeds the array and elements 2 and 3 are both phased −90 degrees, the field at any azimuth θ degrees is:

$$I = 1 /0° + 1 /-90° + 90 \cos \theta° + 51.962 \sin \theta° + 1 /-90° + 90 \cos \theta° - 51.962 \sin \theta°$$

Substituting values for $\theta°$, we get:

$\theta°$	i (relative current magnitude)
0	3.00
30	2.78
60	2.24
90	1.59
120	1.00
150	0.85
180	1.00

Refer to **fig. 7** for a relative power plot of this array.

This graph requires explanation, since I have further manipulated the results for the portrayal of this data. First, the results are normalized, by dividing each result by the maximum value. Second, the logarithm (base ten) is taken of each normalized value and multiplied by 20 to make all the calculated points relative to 0 dB. Thus:

$$dB\ (at\ azimuth\ \theta°) = 20 \log_{10} I/I_{max}$$

Since the maximum value for I occurs at $0°$ azimuth, then normalizing to this value in terms of dB for the data listed above:

$\theta°$	decibels
0	+0
30	−0.65
60	−2.55
90	−5.53
120	−9.54
150	−10.99
180	−9.54

This method of representation best displays array rejection capabilities, not easily shown in a polar plot. For example, assume an array with a respectable −30 dB F/B ratio. Whatever scale is used for the direction of maximum signal must now be divided by 1000 to show this rejection. This will appear as little more than a flyspeck on a polar plot and provide no clear indication of variation with azimuth. Suppose we are listening to an S 9 + 30 dB signal at the front of our array; switching the array around, to the rear

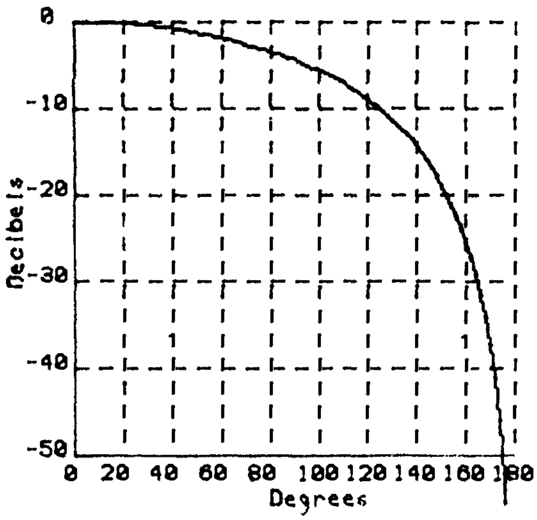

fig. 9. Three-element, triangular configuration, 0.289 wave spacing −90 degree and −90 degree phasing, 1:0.5:0.5 current ratios.

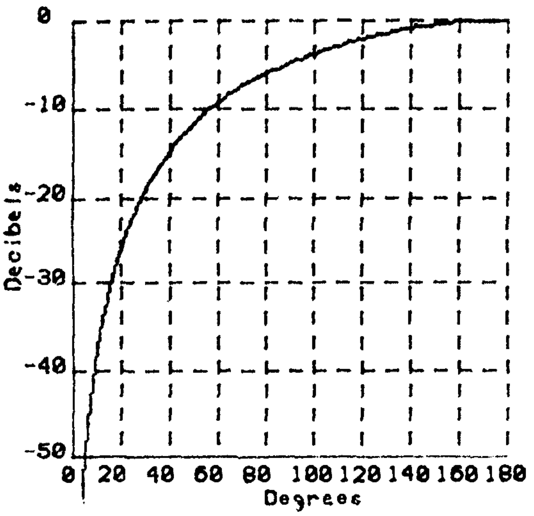

fig. 10. Three-element, triangular configuration, +90 degree and +90 degree phasing, 1:0.5:0.5 current ratios.

we will still see S 9, a not insignificant signal. Yet the reduction is by a factor of *1000*, and *if the transmitter* is running a kilowatt, our array will treat it as though it were only one watt! This illustrates the need for working with logarithmic decibels. But we should not forget what they represent; they are not *linear.*

If the array is symmetrical and has been located symmetrically about the x or y axis, it is not necessary to plot more than 180 degrees; the other half is a mirror image.

determining array gain

The value for *I*, in the direction of maximum signal, is not an *absolute* gain figure. This value is merely derived from the number of elements and the absolute current ratios used. An indication of gain is obtained by observing the total included angle of the main lobe between the half-power (−3 dB) points; thus the smaller the included angle the higher the gain. The simplest way to determine gain is to make a polar power plot (square each azimuth calculation result). Calculate the area of this lobe and then determine the equivalent radius of a circle having the same area. Using the same scale, the ratio of the length of the maximum lobe vector to this equivalent radius is the gain of the array over a single vertical element. However, on the low bands F/B ratio is much more important than gain. For the purposes of making vertical array evaluations from the field equation, keep in mind these implied assumptions:

1. There is an infinitely conducting groundplane.
2. All elements are electrically identical.

Any departure from an infinitely conductive groundplane results in lower efficiency due to ground losses and a higher vertical radiation angle (of maximum signal). If the elements are not electrically identical, the real field pattern differs from the calculated one. For switchable arrays using the same feed network, further complications occur. Even the real field patterns will not be alike.

n-element array calculations

Using the field plotting equation and a programmable calculator, any array layout can be examined. Simply choose the angular coordinates for each element, their drive current amplitude ratios and phase displacements. There are no restraints in choices of current amplitude ratios and phase displacements. (Later, these values will be used in calculating the element driving-point impedances, which in turn will determine the feed network.)

Experimentation shows that equal current to all elements is not always best; neither are element spacings of 1/4 wavelength or current phase displacements of 90 degrees always optimum. **Fig. 11**, an equilateral triangle array, best illustrates these points. This array has elements spaced 1/8 wavelength apart with two of its elements operated at a current amplitude ratio of 0.5 and current phase dis-

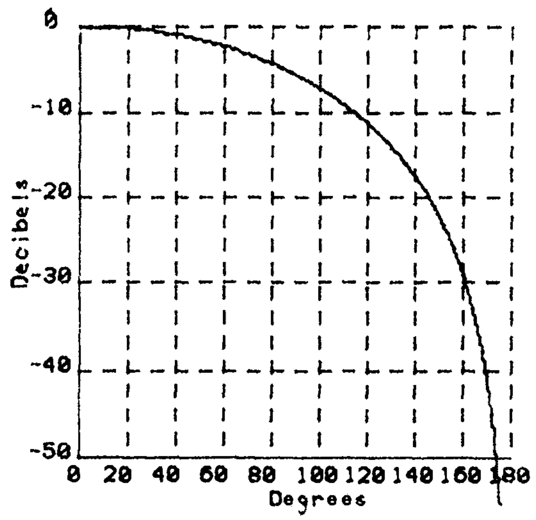

fig. 11. Three-element, triangular configuration, 0.144-wave spacing −135 degree and −135 degree phasing, 1:0.5:0.5 current ratios.

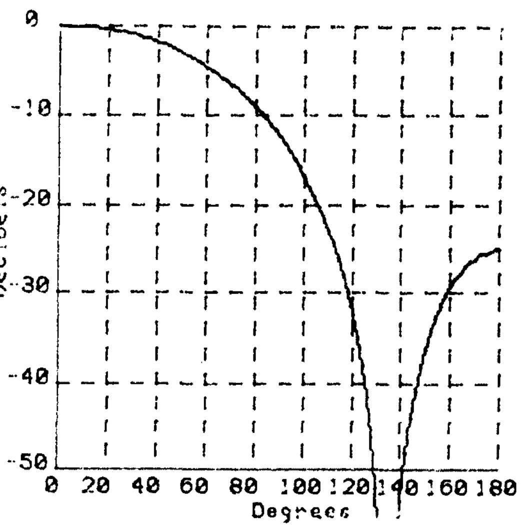

fig. 12. Four-element, 4-square configuration, 1/4-wave spacing −90 degree and −180 degree phasing, 1:1:1:1 current ratios.

placement of −135 degrees. A number of representative arrays have been plotted to show their general properties and to illustrate the variations that occur with changes in physical layout or varying input drive conditions.

two-element arrays

Figs. 3 and 4 are two-element array plots. These produce cardioid field patterns when driven with equal amplitude current with a phase displacement of −90 degrees. The half-power beamwidth is about 180 degrees with a theoretically infinite F/B at precisely 180 degrees azimuth. The 1/8 wavelength spaced array has a slight edge in F/B performance but because of close spacing it has some special problems of its own which I will discuss presently.

The 1/4-wavelength version is quite tolerant of drive condition deviations and displays a useful F/B ratio even with phase displacements far from optimum. These characteristics, plus its simplicity and small space requirement, account for its popularity. The easy tolerance of this design may also account for the unwarranted but widespread belief that more complex arrays will be equally amenable. It so happens that if the feed network of this array consists of 50-ohm, 1/4-wavelength coaxial feeders to the elements and a 1/4-wavelength delay line, nearly optimum drive conditions will exist for best F/B.

triangular arrays

The triangular array plots aptly illustrate design parameter variation (figs. 7, 8, and 9). Prior articles[3,4] proposed 0.289-wavelength element spacing (resulting in 0.25 wavelength for the distance from apex to base of the triangle). W1CF proposed unity current ratios, with two of the elements phased −90 degrees (fig. 7). W2PV proposed phasing these two elements at −110 degrees and reducing their current amplitude ratios to 0.5 (fig. 8). The result for both of these arrays, while providing three alternatives for beam direction, is a not very spectacular maximum F/B ratio of −10 and −15 dB, respectively. If parts of both proposals are combined, that is, phases of −90 degrees and current ratios of 0.5 for these two elements, maximum F/B ratio improves to nearly infinity (fig. 9). A little time spent with a calculator has changed a not-too-interesting array into an exciting performer.

Both writers proposed to make these arrays switchable in six directions; W2PV omitted explaining how, and W1CF proposed an equal power divider. However, since the elements will not present equal and resistive driving-point impedances (in any of these variations), W1CF's intended field plot cannot be achieved with equal power division. Since the half-power beamwidth of these arrays is about 135 degrees, the additional complexity required to switch

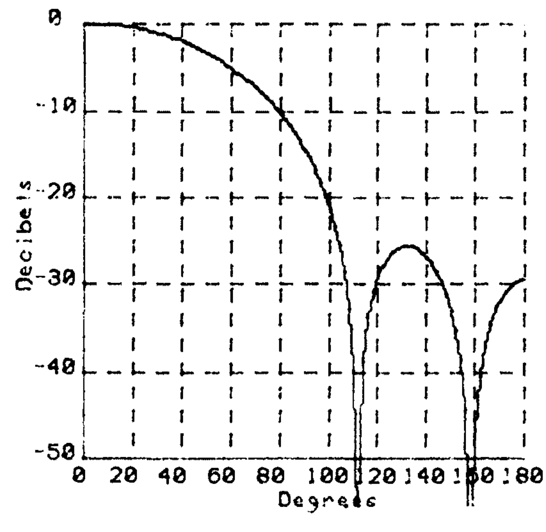

fig. 13. Four-element, 4-square configuration, 0.272-wave spacing −90 degree and −180 degree phasing, 1:1:1:1 current ratios.

this array in six directions makes it of questionable value. Nevertheless, we can develop a feed network which produces a leading phase of +90 degrees, making the array switchable in six directions (see **fig. 10**). Two feed networks are required. A future article will present a lumped constant network equivalent to coaxial lines, except that current phase may be advanced as well as delayed (and may be designed for any characteristic impedance one happens to require!).

three-element in-line arrays

This antenna is another example of an unequal current-amplitude-ratio driven array. The middle element *current* is twice that of the reference element. The 1/4-wavelength spaced version, properly driven, has a 90 degree included angle over which the F/B is better than −25 dB (approaching infinity at the 180 degree azimuth). The half-power beamwidth is about 150 degrees and is down −6 dB, or one S-unit, at the ±90 degree azimuths. On 80 meters the F/B capability of this array has been impressive in listening tests, even with nearby stations (within 20 miles), the ultimate test of F/B on the low bands. I have often considered the possibilities of an antenna consisting of two such arrays, operated at right angles to each other.

The 1/8-wavelength spaced array plot is a near duplicate of the wider spaced array. It has slightly higher gain as a result of its narrower half-power beamwidth of 110 degrees and it has a wider width over which F/B exceeds −25 dB.

4-square arrays

These arrays, although having four elements, are closely related to the three-element in-line type. The array projects its main lobe along a diagonal of the square, with the two middle elements driven at the same phase and the current divided between them. In effect, the middle element is split into two elements. Half-power beamwidth is about 95 degrees, indicating a gain increase over the three-element inline. The width over which the F/B is −25 dB or better has increased to 150 degrees, though the average rejection over this range is not as deep as with three-element in-line arrays. The symmetry of this element arrangement allows the array to be switched in four directions using the same feed network; that is, the main lobe may be formed in either direction along either diagonal. As has been pointed out earlier, due to the significant dissimilarity of drive-point impedances of any element as the array direction is switched, more than ordinary care must be taken to ensure electrical symmetry. Experiments with the field equation demonstrates the high minor lobe sensitivity of this array to small deviations in any of its design parameters. For example, changing element spacing from 0.25 wavelength (**fig. 12**) to 0.272 results in the formation of two additional minor lobes at

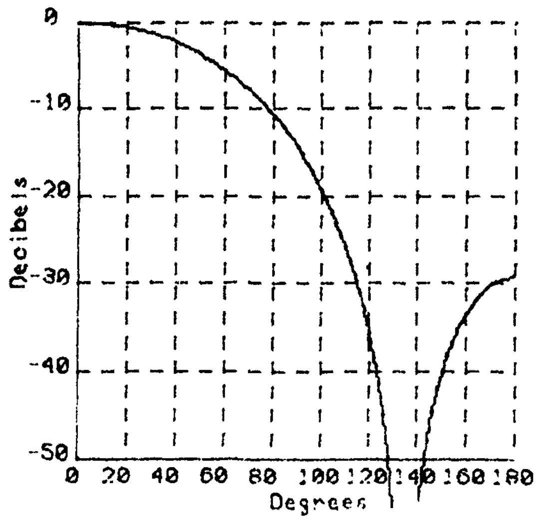

fig. 14. Four-element, 4-square configuration, 1/8-wave spacing −135 degree and −270 degree phasing, 1:1:1:1 current ratios.

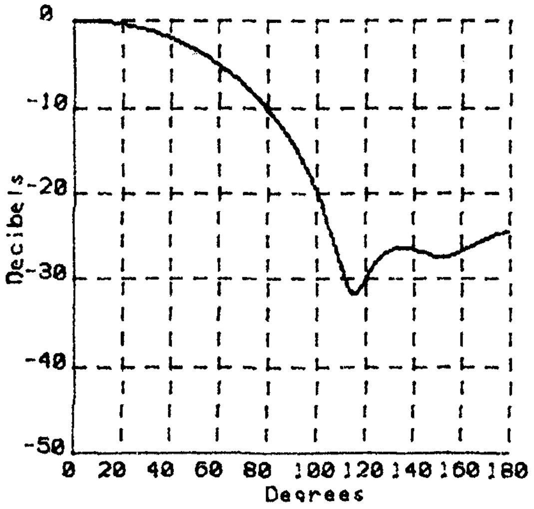

fig. 15. Four-element, 4-square configuration, 1/4-wave spacing −90 degree and −190 degree phasing, 1:1:1:1 current ratios.

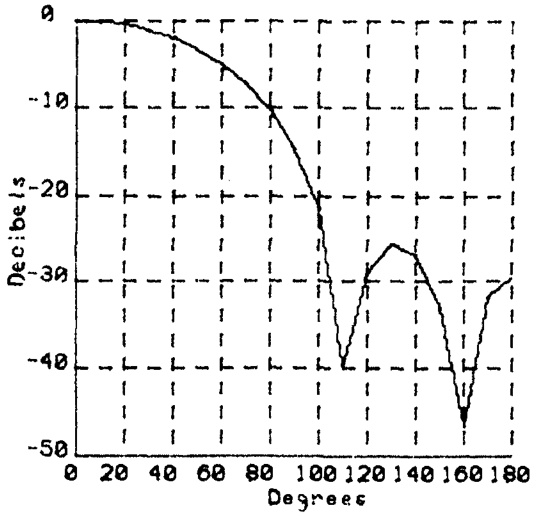

fig. 16. Four-element, 4-square configuration, 0.272-wave spacing −90 degree and −180 degree phasing, 1:1:1:1 current ratios. Poor granularity (plotted in steps which are too large).

130 and 230 degrees (fig. 13). If the phasing is changed from exact multiples to −90 and −190 degrees (fig. 15), the additional lobes have formed but without definition to the nulls. The same sensitivity is shown to small dissimilarities in drive current ratios among the elements.

array of arrays

This name refers to an antenna arranged to consist of arrays which are themselves arrayed. The simplest example consists of two two-element arrays configured as a square. Two adjacent elements are treated as reference elements fed in phase, and the remaining two elements are also fed in phase but displaced −90 degrees. Current amplitudes are all equal. This scheme allows switching the main lobe in four directions, except these are offset 45 degrees from the diagonal directions. In combination with the 4 square feed network we could have eight directions. As commented in connection with the triangular array, since the half-power beamwidth of the 4-square fed connection is about 95 degrees, switching in a different feed network for these additional directions appears to be a needless complication. (Perhaps more useful would be the provision for a separate feed network for optimum F/B operation of the array in the 80-meter phone and CW subbands.)

With 1/4 wave spacing, −90 degrees phasing and all elements fed equal current (amplitude), the half-power beamwidth is about 140 degrees as seen in fig. 7. The −25 dB or better F/B width is a paltry 40 degrees, although at 180 degrees azimuth it approaches infinity. Except for increased gain over a single two-element array, the performance of this antenna is not notable and is not nearly equivalent to that obtained from the same physical layout when connected as a 4-square.

As noted earlier for simpler arrays, the 1/8-wavelength-spaced 4-square field pattern is nearly identical to the 1/4-wavelength-spaced pattern. In each type of array examined we can note differences which show improvements in all characteristics over its equivalent larger size array. However, closer spacing means high mutual coupling, which in turn means even greater sensitivity to element variations. Such an array is difficult to provide identical field patterns for all switchable directions. Unless you have prior experience with these arrays and have equipment for accurate measurements of the self- and mutual impedances of the array elements, you are strongly advised to avoid these closely spaced arrays.

One-eighth-wavelength arrays present an interesting challenge and some opportunities. They offer the same array in much less area — and two-band operation is possible. But if the height of the elements is 1/8 wavelength at the lowest frequency, the self-im-

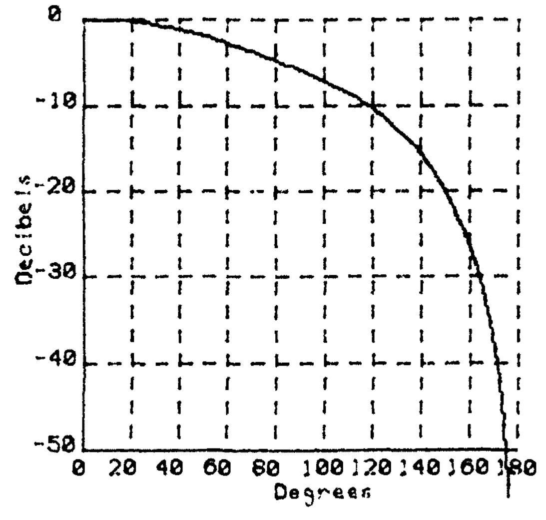

fig. 17. Four-element, twin two-element array, arranged in a square configuration, 1/4-wave spacing −90 and −90 degree phasing, 1:1:1:1 current ratios. This array of arrays does *not* use the 4-square feed network.

pedances are going to be quite low. The resistive component will be about 6 ohms — not easy to work with, and placing a high premium on the need for a low-loss groundplane. If the element length is significantly more than 1/4 wavelength at the highest frequency, temporary sectioning of these elements has to be provided so that impedance measurements can be made (the elements have to be electrically separable into 1/4 wavelength or shorter sections for the measurements).

other possibilities

An interesting possibility is a five-element array which places an additional element in the center of a 4-square. Since this element is always occupying the same electrical position in the array regardless of beam direction it represents no increased switching complication. F/B ratio width can be increased over that of the 4-square even if −30 dB is used as the limiting criterion. Alternatively, if a way could be found to keep the "outrigger" elements from entering into the act, this arrangement of elements could also be operated as crossed three-element in-line arrays. Although there would be some loss in gain, the tradeoff is a significant improvement in F/B depth.

conclusion

So much for the theoretical design. With the concepts and suggestions reviewed in this article, I hope I have given experimenters the tools and some ideas for selecting and siting an array.

The next part of this series deals with self- and mutual impedances; how to measure them, and, most crucial of all, what to do with them. Until we know the driving-point impedances, the feed network design cannot proceed.

acknowledgement

I am indebted to WB6SXV who derived the field strength algorithm used.

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ham radio

smart squelch

A circuit to enhance voice reception in the presence of interference

I did not plan to build a squelch. But now that the design is complete, I cannot find a better description. What I started out to do was simply to automate a feature that is already available on my Comm Audio Processor (CAP), Model 210.[1] I have named this new device the smart squelch or SSB squelch, because it is considerably more involved than those simple systems that just open an audio channel when signal energy becomes present in a receiver's i-f amplifier.

Most of those who bought or built the CAP design soon discovered that tuning an SSB signal is easier if you reduce gain on the high-band side of the binaural audio pair and then tune until vowel and vowel-like sounds fill the low-band channel. Then, you just raise gain on the high-band again to obtain full comprehension and clarity. The explanation is that, with SSB, we are forced to tune an audio bandwidth at rf frequencies across a fixed oscillator associated with a receiver's product detector until the heterodyne process produces audio frequencies that sound normal. Tuning from one side you first get that high-pitched, nasal-sounding monkey chatter as the powerful, and relatively long, voiced energy is heterodyned to produce higher than normal audio frequencies. Conversely, tuning from the other side first produces sort of "wush, wush" sounds, as shorter high-frequency transient-like sounds translate at lower than normal audio frequencies. Of the two, the high-frequency nasal sounds contain much more energy and seem to be the more offensive.

Clearly, it seemed, I would have to detect only when the low-band binaural audio band contains signal energy, and use that information to *un*-squelch a normally squelched high-band channel. This would automate the rejection of the offensive monkey chatter, and in the process make tuning easier. Sounds simple, it is, and it works — but only when signal-to-noise ratio is good and noise interference is minimal. I wanted a system that would work in the real world of interference and noise.

My design solution employs a particular set of filters and a non-linear detection process that regenerates the fundamental frequency of laryngeal voice sounds. This frequency is then rectified and used as a control signal. With this technique, much of the energy that would often false-trigger the squelch is rejected in the detection system. In effect, the squelch system is matched to particular aspects of human speech. Understanding my system requires some understanding of how humans speak.

human speech

Fig. 1 is a drawing made to simulate a voice spectrogram. To produce such a spectrogram, a phrase one or two seconds long is recorded and then repeated to give a picture showing the frequencies generated and a crude indication of relative amplitudes. Key features to note include the relative frequencies, the duration of various parts of speech, and the typical time between utterances.

The fricative and plosive parts of speech contain energy starting from 1 or 2 kHz and rising to as high as 10 kHz or more. In our typical receiver's 3-kHz audio bandwidth we throw away most of this energy without too much loss in comprehension. But without some of this energy, however, words such as *sat* and *fat* would simply be heard as *at*.

Voiced energy appears in lower-frequency regions of the sound spectrum (voiced speech means sounds

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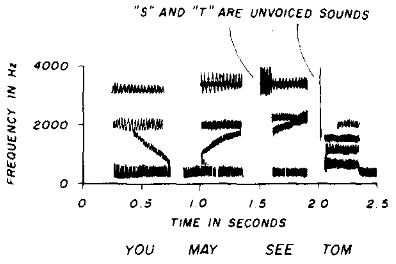

fig. 1. Spectrogram of the sentence "You may see Tom."

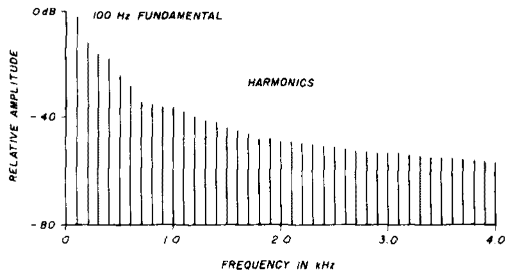

fig. 2. Simulated output of the human larynx.

produced by vibrating the vocal cords, such as in the enunciation of the vowels a, e, i, o, u, or the nasals m or n; unvoiced sounds, such as the consonants *t* and *p*, are produced without the vocal cords). To get a feel for how voiced sounds are made, consider what happens when you fill a toy balloon with air and then let the air escape through stretched lips at the end of the neck (this simulates our lungs, vocal cords, and larynx). The sound emitted is similar to the sound that comes from our own vocal cords.

Fig. 2 shows a spectrum of sound recorded right at the vocal cords. **Fig. 3** shows what comes out (that is, what we actually hear) when the spectrum of **fig. 2** is filtered through the acoustic chambers of our mouth and nasal passages. The areas of relative peak amplitude are called formants. And where the formants are located is determined by how you hold your mouth while sounding your vocal cords.

Fig. 3 shows the detailed structure and the relatively accurate amplitudes for a short time interval in the spectogram; the abscissa of **fig. 3** corresponds to the ordinate of **fig. 1** at a point in time when the formants match. In man, the voiced fundamental frequency, indicated by the first line and separation between all harmonic lines, lies mostly between 80 and 160 Hz. The first — and most powerful — formant ranges from about 300 to 600 Hz. These frequencies, and an average 0.3-second gap between elements of speech (observed as average in many spectrograms) provide the key factors to the system's design.

References 2 through 5 provide more information on speech and its production.

system requirements

The block diagram in **fig. 4** shows everything required to add the squelch system to an existing receiver. If you already have a CAP or something similar, you need only the laryngeal fundamental detector, timing control circuits, and solid-state switch.

Since the binaural synthesizer has been covered in previous articles, a functional description will not be repeated. Note, however, that the binaural crossover frequency is set at 750 Hz. This choice was made originally because most receivers have chosen this for CW band center. A very slight improvement would theoretically result if 600 Hz were used for the cross-over when using the squelch system. I could not hear an audible improvement, however, when 600 Hz was tried, so the cross-over was left at 750 Hz. Performance was quite good, however, when the squelch system was used with the CAP's **VOICE**

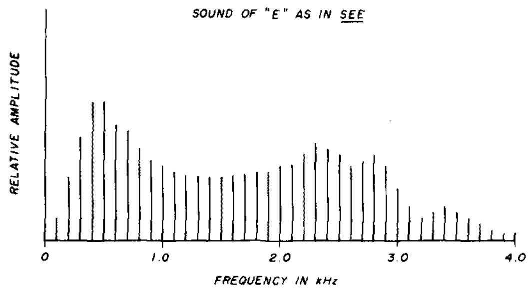

fig. 3. Simulated vocal tract output.

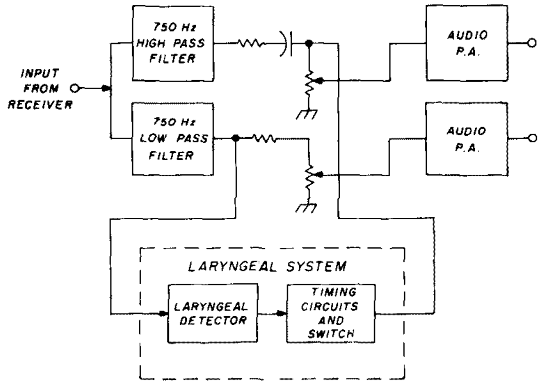

fig. 4. The basic squelch system.

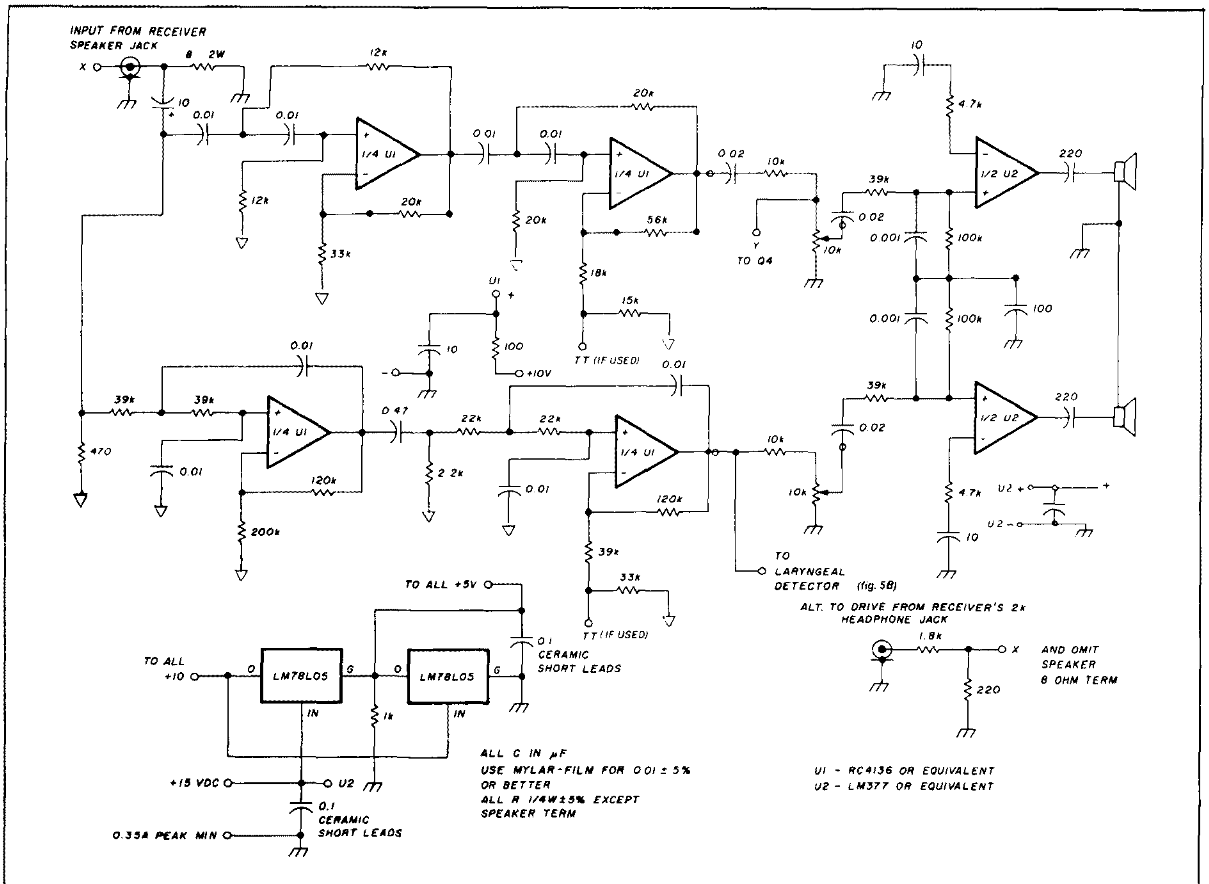

fig. 5A. The laryngeal detector.

FILTER position. A description of the remaining circuits follows.

laryngeal fundamental detector

A level sensitivity control precedes the first filter. That filter is designed to accept and amplify two or more of the voiced fundamental harmonics in the general frequency range of the first formant. These harmonics are separated by an amount equal to the fundamental frequency, and feeding a pair or more of them into a diode and load circuit will result in the generation of a large number of harmonics — as well as sum and difference frequencies. The second filter/amplifier selects the difference frequency and rejects all others, effectively reconstituting the fundamental voice frequency without any of its harmonics. This sine-wave energy is then fed to an absolute value circuit (a full-wave rectifier).

The result of all this is the fast appearance of a dc level when your receiver is tuned at — or very nearly at — the proper frequency for an incoming SSB signal. Remaining circuitry provides a fast release of switch Q4. This switch keeps the high-band audio channel normally squelched — until tuning presents a normal (or nearly normal) voice signal to the detector. Finally, a selectable "hang-time" is provided to enable the system to re-squelch quickly, or to remain open long enough to bridge time gaps between the words and phrases in normal speech, or to remain open somewhat longer still.

interference rejection

Ideally, the 300 to 600 Hz filter assembly should reject any audio energy outside of its bandwidth. Also, no single-line signal (and no dual-line signal with a frequency separation of more than 160 Hz within the 300 Hz bandwidth) can produce a difference frequency at the diode load output that can get through the 80 to 160 Hz filter assembly. The system is thus matched to basic voice characteristics, making it much better at interference rejection than the first technique. The interference rejection is not perfect, however, because various combinations of other signals and high noise levels will occasionally meet the detection requirements. When this happens the squelch opens, looks around for the hang-time period, and clamps down again.

circuit details

Fig. 5 shows the complete circuit with binaural synthesizer and a 2-watt audio output capability.* A level-setting potentiometer is connected to the low-band audio channel to provide an input to the 300 to 600 Hz filter, which is made up of an active staggered pair bandpass filter driving an active two-pole high-pass filter. The output from this filter is fed to a diode

*Write to Hildreth Engineering, P.O. Box 60003, Sunnyvale, California 94088 for PC board and kit information.

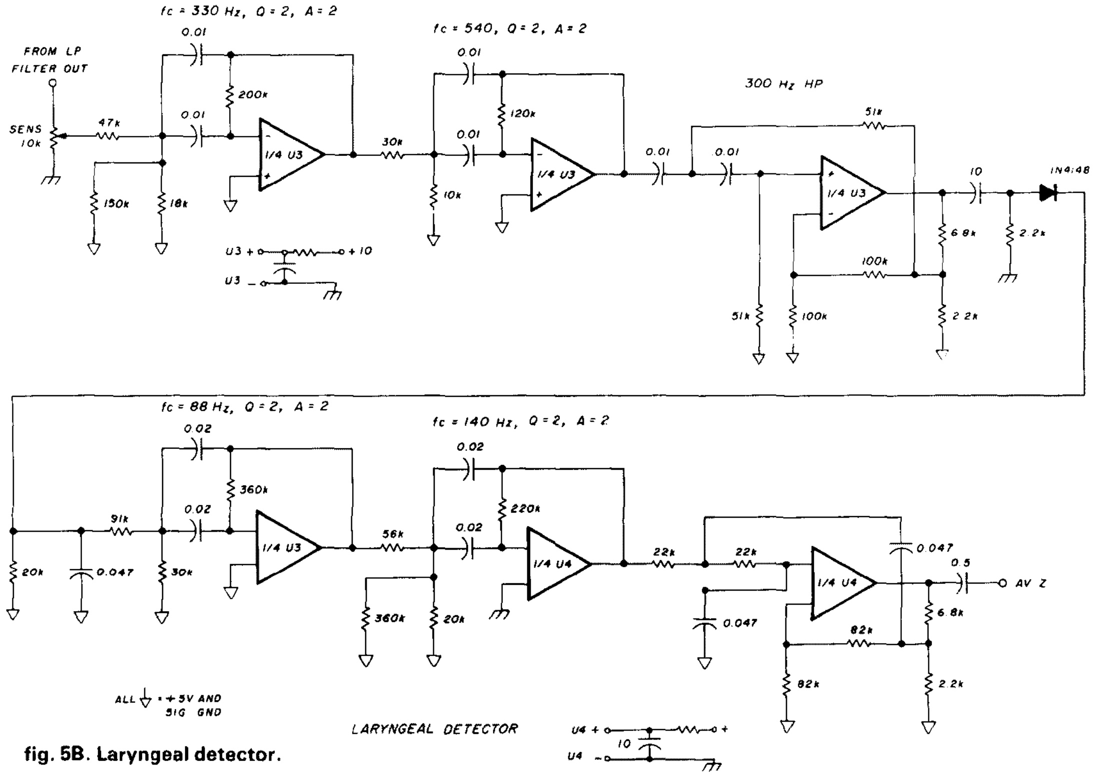

fig. 5B. Laryngeal detector.

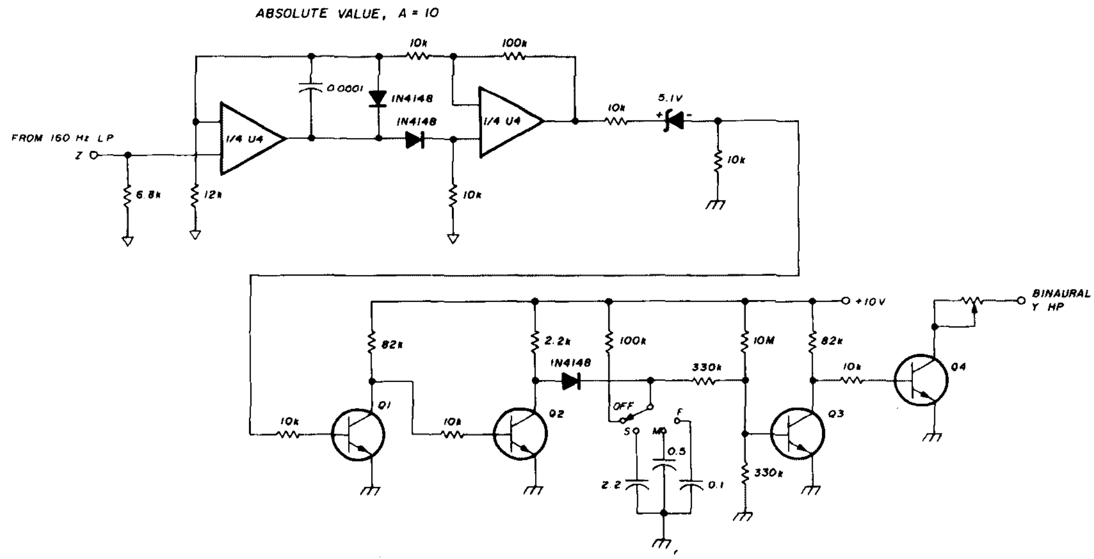

fig. 5B (cont'd.)

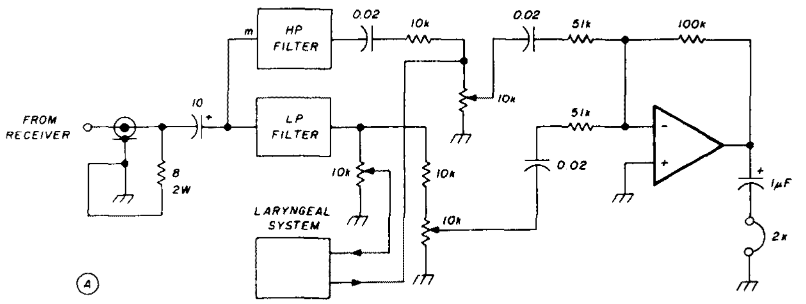

fig. 6A. Combined spectrum for monaural sum to headsets.

and its load circuit. Following the non-linear process is an active, staggered-pair bandpass filter and an active, two-pole lowpass filter that provides an 80 to 160 Hz bandpass with a steep cutoff above 160 Hz. The output from this filter is fed to an absolute value circuit, which also provides an additional voltage gain of 10. Transistors Q1 through Q3 provide start and stop timing control, and Q4 functions as a clamp. Zener diode CR1 restores the no-signal +5 volt input from the absolute value circuit to ground reference, thus ensuring that CR1 is off. This, in turn, results in Q2 and Q4 being on, and Q3 being off.

When a signal activates the laryngeal detector, Q1 conducts and shuts Q2 off. The timing capacitor is quickly charged through RT1, which results in Q3 being on and Q4 being off, thereby *un*-squelching the high-band audio channel. When an input signal disappears Q2 clamps, but the voltage is left high on the timing capacitor, C, which must now discharge through RT2 before Q4 can re-clamp. Three values of C are shown from which you can choose fast, medium, or slow squelch. The last switch position locks the squelch off.

The power supply system uses small, totem-poled, 5-volt regulators with a pull-down resistor to provide +10 volts and +5 volts at signal ground for op-amp bias. This arrangement works well with a +15 volt main source. It will function down to +12.5 volts with some loss in available audio power.

construction details

This system employs a collection of relatively uncritical circuits. All filters are of low Q, which makes the use of 5-percent tolerance resistors and capacitors adequate. Paralleled trimming resistors are used in a few places where the calculated value falls right in between the normal 5-percent value. Calculated center frequency values are given for each bandpass filter — and this is the most critical factor. Center frequency of these elements may be "tweaked," if desired, by adjusting the resistance value on the resistor from the two capacitor and input resistor junction to ground. NPO ceramic capacitors and precision resistors are recommended only if the unit is to be used in extreme environments.

operation

It is best to first turn the squelch off and get used to tuning SSB signals with the normal binaural system, as indicated in the CAP article. Then, once you have a properly tuned signal and the squelch sensitivity control all the way down, switch the squelch control to **FAST**. This will immediately squelch the high-band channel. (The degree of squelch is determined by the resistance between the collector of Q4 and the high-band gain control. With R at 0 ohms you get maximum squelch; at 500 ohms you get 20 dB.) As you bring the sensitivity control up, a point will be reached at which the high-band audio will come on. Crank the control up a little higher to take care of any fade, then just listen for a while. Unless the speaker keeps up a fast monologue without pause, you will get an occasional squelch during pauses in speech, followed by an immediate squelch release as speech continues.

How fast to set the hang-time is a matter of personal preference and band conditions. If you are listening to someone who has the habit of saying, "Well . . . (3 or 4 seconds) . . . I talked to Joe the other night . . . .," you may want the squelch on fast if there is nearby interference, so that you won't be bothered with that other stuff during the hesitation. On the other hand, if conditions are quiet, you may want the squelch on slow for less chop. A careful listener will note that the squelch system will shave unvoiced sound if it is at the beginning of an opening statement. As it turns out, context saves the day and you just won't miss that occasional burst of noise.

Other variables include relative gain settings on the binaural channels as well as your receiver's AGC time constant, and **OFF** selection in addition to the relative settings of your rf and af gain controls. No doubt there are more.

✓ 174

✓ 135

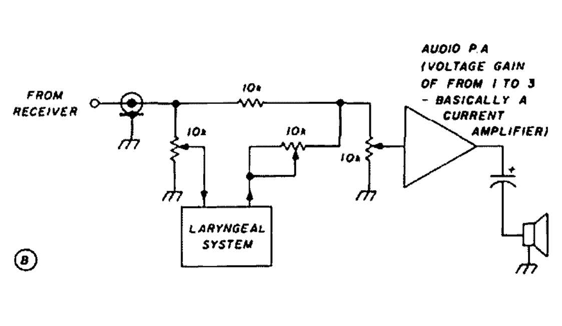

fig. 6B. A simple system, useful in certain applications.

alternative applications

Fig. 6 shows two of many potential applications. Those who use headsets often prefer the same spectrum in both ears. This is provided by using the circuit in **fig. 6A**. You can also get this with speakers by making them coaxial. The simple application in **fig. 6B** omits the SSB tuning assistance advantage, but may be useful in some cases.

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glossary

low-band — The audio band from 100 Hz nominal to 750 Hz.
high-band — The audio band from 750 Hz to a nominal 3 kHz.
laryngeal — Sounds produced with the human larynx.
fricative — Speech sound generated by breath forced through a restricted area, such as the sounds s and v.
plosive — Speech sounds created when the breath is expelled suddenly from a completely closed oral cavity, as in the sounds of p and d.
formant — Regions of a sound spectrum in which certain of the harmonics of laryngeal sound are passed with minimal attenuation (that is, areas of greatest amplitude). They are seen in the dark bands of **fig. 1** and the peaked areas of **fig. 3**.
hang-time — The time period during which the squelch remains open after voice sounds stop.
timing management — Control of rise, fall, or delay times.
phoneme — A member of the set of the smallest units of speech that serve to distinguish one utterance from another in a language or dialect.

ham radio

harmonic product detector
for QRP transceivers

A circuit that overcomes amplifier feedback problems in small QRP rigs

Compact QRP transceivers that use a direct-conversion receiver on **receive** and a chain of amplifiers to boost a VFO signal in **transmit** often produce poor-quality transmitted signals. This is generally caused by insufficient decoupling or shielding in the amplifier chain, which permits feedback from one or more of the amplifiers to get into the VFO. That results in instability in the form of frequency pulling, rough note, and the like.

I first noticed this problem during the construction of a QRP transceiver similar to the unit described by W7EL.[1] Although my unit was not quite as compact as his it was still crowded, because my version incorporated two bands, a keyer, and two 0.9-AH rechargeable batteries all in a 4 × 6 × 2-inch (10.16 × 15.24 × 5.08 cm) box. Naturally, the smaller the rig the worse the problem, because of the proximity of components and wiring.

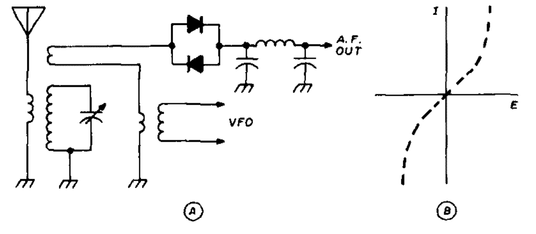

fig. 1. (A) Basic back-to-back mixer. (B) Voltage-current characteristics of the back-to-back matched diodes.

While trying to think of a solution to this problem, I recalled the old ECO (electron-coupled oscillator) of many years ago. The approach there was to use a screen grid tube like the 6F6. The grid circuit operated at one-half the desired output frequency, and the plate circuit doubled to the output frequency. Even with the wide-open breadboard layout used in those days, the transmitted signal was quite acceptable. Feedback into the ECO grid circuit had little or no effect on signal quality because of the difference in frequency between the grid and plate circuits and that of the following stages.

Adapting this concept to my two-band solid-state transceiver (without adding more stages) became a possibility when I discovered in the literature a unique product detector that requires oscillator injection at one-half the output frequency, thereby minimizing the feedback problem.[2]

The basic harmonic back-to-back mixer (or detector) is shown in **fig. 1A**. As explained by the author, V. Polyakov, RA3AEE, the diodes provide the voltage-current characteristic shown by the dotted line in **fig. 1B**. It is a symmetrical, cubical parabola, which can be achieved by connecting two matched diodes back-to-back.

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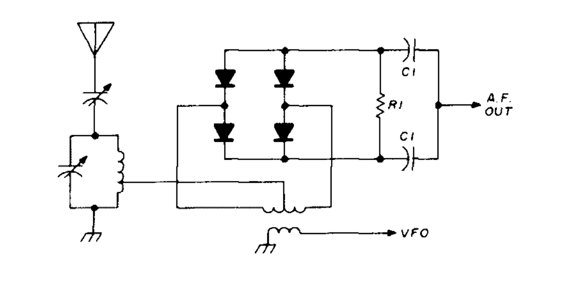

fig. 2. Improved mixer using a balanced arrangement plus bias via R1 and C1.

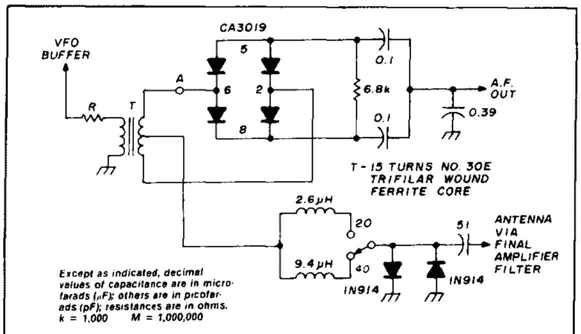

fig. 3. Harmonic mixer circuit used in transceiver.

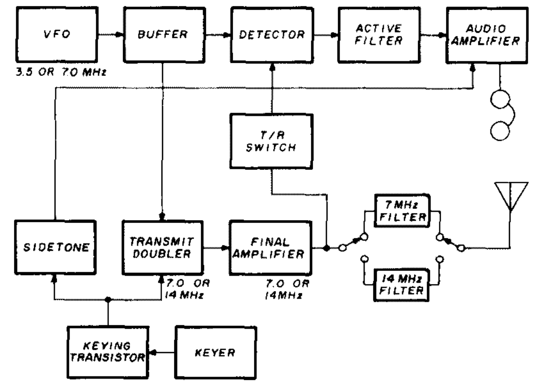

fig. 4. Block diagram of QRP transceiver, based on the W7EL design, and using the harmonic product detector.

RA3AEE describes the operation of the detector as follows: "When the local oscillator voltage goes through zero, both diodes are open circuit and the circuit current vanishes. At the peaks of both positive and negative half-waves of this voltage, one or the other of the diodes conducts and the signal source is connected to the load. In this way the mixer works like a switch, closing the circuit at a frequency equal to twice that of the local oscillator."

This mixer has two significant characteristics. First, the local oscillator must be tuned to a frequency one-half that of the incoming signal. Second, there is no direct current in the load circuit, which means that signals from high-power interfering stations are not detected and thus produce no noise. This second characteristic is true only if symmetry is preserved in the detector.

The circuit has several disadvantages, namely loss of signal power in coupling to the local oscillator and excessive sensitivity to oscillator injection for optimum conversion gain. RA3AAE devised several revisions to overcome these problems (see **fig. 2**). A balanced arrangement using four diodes overcomes the coupling loss, and, by applying a bias voltage to the diodes via R1 C1, the circuit becomes less sensitive to oscillator injection.

Fig. 3 shows the final version of this detector, the one I used. Oscillator injection at point **A** should be adjusted by means of resistor R to achieve approximately 1 volt RMS. Too low an oscillator injection will reduce conversion gain, but I noticed no ill effects from higher levels of injection. The optimum value of R in my case was 270 ohms, but this value will depend on individual circuitry.

I used an RCA CA3019 diode IC for the detector with pin connections as shown. The usual 1N914 or hot carrier diodes would also be suitable. The RA3AAE version used germanium diodes, and the reported optimum oscillator injection voltage for these was 0.6 to 0.7 volts RMS.

My transceiver follows the basic pattern of the W7EL unit, shown in block form in **fig. 4**. The driver transistor, however, now becomes a doubler. There remains enough drive for the desired input of 2 watts to the final, for output on 40 or 20 meters of about 1 watt.

I have a 5-kW broadcast station almost in my back yard, and I was pleased to find that the harmonic detector, as claimed, was not susceptible to fundamental overload. The original article also claims excellent isolation between oscillator and antenna. Finally, because the oscillator is operating at a lower frequency, drift is reduced and stability enhanced.

For those interested in the higher frequencies, see the article by WAØRDX on the twin-diode microwave mixer in *ham radio*, October, 1978, page 84.

references

1. "An Optimized QRP Transceiver," *QST*, August, 1980, page
2. *Amateur Radio Techniques*, Seventh Edition, Radio Society of Great Britain, pages 131 and 132.

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

The technical advances made in ham radio during the last two decades have been amazing. Moonbounce (EME) and satellite communications are now commonplace. Some avid operators have made DXCC via satellite and others have made WAC via moonbounce.

On the lower frequency bands, the technical advances have been equally impressive. Solid state has supplanted vacuum tubes up to power levels of several hundred watts and a new order of frequency control has been provided by digitized, frequency-synthesized transceivers using a high-stability crystal oscillator as master frequency control. Frequency synthesis with 10-Hz readout is practical for ham equipment and synthesis to 1 Hz is often used in commercial gear.

So here we are in mid 1983, with all this great equipment that would make the Amateur of the sixties turn green with envy. But what price have we paid for these technical advances?

One problem with the new gear that's showing up on the ham bands, but which has been largely ignored except by those affected, is that of white noise interference.

White noise is uniform-spectrum, random noise heard as an unwanted signal in communications equipment. It is caused primarily by random motion of electrons in a circuit and has been a well-known problem in telephone circuits for years. It's been largely overlooked in Amateur Radio because the signal levels encountered in the past were great enough so that white noise was not a problem — and also because the relatively unsophisticated circuits used until recently did not generate much white noise. That is no longer true. Improvement in receiver sensitivity, stability, and dynamic range, coupled with frequency synthesis and phase-locked-loop systems, has created a white noise problem that is hard to ignore today. Here's a typical example of a white noise problem that was encountered by a good friend of mine:

My friend Joe finally traded in his VFO-controlled, tube-type transceiver for a brand new, solid-state, synthesized PLL rig. He was hugely pleased with its sensitivity, stability, dynamic range, and its frequency readout to 100 Hz. When his close friend John (who lives on the next block) decided to get a new rig, Joe encouraged him to buy a transceiver just like his. John did, and soon there were two identical transceivers on the air in the same neighborhood.

Up till then the two friends had operated with no problems. Each ran about 200 watts input to a high-grade, tube-type transceiver, and when their triband beams were not pointed directly at each other, causing receiver overload, they could both work the same band at the same time. When they worked different bands, neither knew the other was on the air. This situation had existed for many years. But soon both Amateurs were dismayed to discover that their brand new transceivers generated so much white noise hash that operation on the same band at the same time was impossible. Even operation on *adjacent bands* was marred by an S3 background noise level from the nearby transceiver!

Joe told me, "When my friend is listening on the same band I am listening to, I hear a rushing noise within 30 kHz of where he is tuning. When he transmits, the white noise completely wipes out reception on that band and ruins reception on the adjacent bands. And my transceiver does the same to him!"

The white noise problem is plaguing another local Amateur: Harry lives

about a quarter mile from another DXer. Both Amateurs run full power. Harry has a tube-type receiver and transmitter of the highest quality. His friend has a solid-state receiver and exciter. Harry tells me his reception is completely blocked out when he aims his beam at the other DXer's house. Unfortunately, that direction is the direct path to Europe! When both are operating at the "low end" of 20 meters, most DX signals below S5 or thereabouts are lost in the white noise hiss generated by the solid-state equipment hooked up to the nearby DXer's beam. For the last six months Harry hasn't had many European contacts, and he is *very* unhappy about it.

Disturbed by these stories, I called up a local ham who has a solid-state, synthesized hf transceiver. I ran some tests, listening to his equipment while he was both receiving and transmitting. The results were negative: I heard no white noise at all from his rig, which was about two miles away.

After checking with some other Amateurs in the area I came to the reluctant conclusions that little is known about the white noise problem with respect to Amateur gear, that the amount of noise interference generated by hf gear varies between manufacturers from type to type, and even from unit to unit. One rig may generate a lot of white noise interference for nearby Amateurs while another will be relatively clear of it. No general conclusions can be drawn about individual pieces of equipment at this time, except that the problem seems to be widespread.

Does your equipment generate white noise interference? Ask a neighborhood ham to listen on the skirts of your signal for the characteristic hissing sound. It may go out as far as 100 kHz from the carrier frequency. Or, if you have an auxiliary receiver, remove the antenna and listen yourself. You may be surprised!

I'll be interested to hear from any Amateurs on this subject. Let me

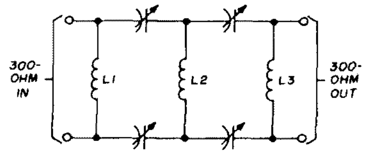

fig. 1. The K5BDZ TVI filter. The capacitors are 20-pF variable ceramic units. Coils L1 and L3 are forty turns No. 30, closewound, 1/8-inch diameter. Coil L2 is twenty-two turns No. 30 closewound, 1/8-inch diameter.

know the type of equipment and the noise problem you encounter.

I think the subject of white noise is best summed up by an editorial in the newsletter *Amateur Radio Today* which says, in part, "The question is not whether the hash exists, but rather to what degree, and whether or not it is bothersome . . .

"While *any* unnecessary hash in a receiver is unwanted, we at *Amateur Radio Today* feel that the tradeoff between a small amount of noise and having the frequency-selection flexibility offered by a synthesizer may be worthwhile. That's a judgment each individual will have to make."

a thoughtful review of signal reporting

The present RST system for reporting signal strength was devised by Art Braaten, W2BSR, about 1934. It certainly has withstood the test of time; it is still being used nearly fifty years later. Old Timers will remember the previous QSA and R system, (Your sigs are QSA 5 R9), which rapidly faded away as the RST system took over.

It is apparent, however, that flaws are appearing in the venerable RST system. Sideband operators have expanded the system to include "decibels over S9" and CW operators have clipped the system so an RST 599 signal is now 5NN. Obviously, the technical advances in communications over the last few decades have outstripped the system of reporting signal strength.

two solutions to the RST problem

It is simple enough to add decibel ratings at the top end of the scale as the sidebanders have done. CW operators could use the scheme by substituting an exclamation point for each additional 10 dB of signal strength. Example: 5NN!! means 20 dB over 599. Simple!

But the system really breaks down when it comes to weak-signal reporting. Nobody wants to give an S1 or an S0 report, and some rare DX signals are really weaker than that!

Two 1983-style solutions to the weak signal report have surfaced recently. One system is for weak-signal (moonbounce) VHF work, and the other for 75-meter DX work. Both bands are well known for long-distance DX contacts under adverse conditions.

The problems seem to arise when the DX signal is virtually buried in the noise. Listening to such a signal after a length of time can produce some queer effects on the listener. He imagines that he hears the DX signal calling him! Perhaps it is only an uneven change in the background noise, or merely a hunch, but the listener is *sure* he is in QSO with the almost-inaudible DX station. And, by George, many times he is correct.

the 80-meter reporting system

Experienced DXers are well aware of this phenomenon. (In fact, some DXers can work a station when it isn't even there. But such a situation is outside the scope of this discussion.)

The 80-meter reporting system (attributable to Gordon, W7FU, I believe) is termed the "ESP System," the initials standing for extra-sensory perception. Signals are graded on an ESP scale of one to five. The scale

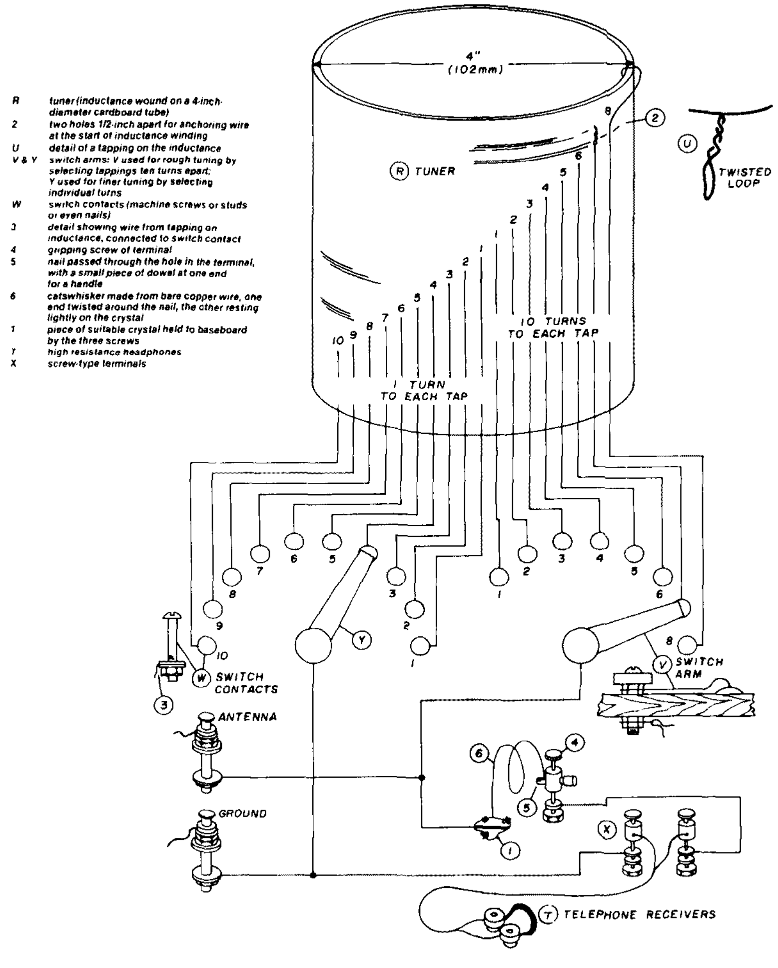

fig. 2. Construction details of the U.S. Bureau of Standards crystal radio receiver (1920).

has not been standardized, as far as I know, the outer limits being ESP-1 ("the DX station is thinking of coming on the air,") to ESP-5 ("the DX station is inaudible but workable"). From all reports, the ESP system seems to be functioning quite well for both sideband and CW service on the lower frequency bands. I'll be happy to report on this exciting new aspect of Amateur Radio as it develops.

the EME (moonbounce) reporting system

I understand this signal reporting system was developed by Dave, K1WHS. During low-signal moonbounce work he noticed that very weak signals could be described as "musical noise," that is, noise that varied in tone as the receiver tuning dial was moved slightly. After listening to an evening of such noise, trying to dig a signal out of the background that he knew was there, Dave coined the term "Imagination-Enhanced QSO," the signal strength of which could be reported in IEQ units ranging from one to five. The IEQ scale has not been established yet, as most IEQ reports have run between zero and one. Much work remains to be done on this interesting system of signal reporting, however, before it becomes practical. Stay tuned in and I'll keep you up to date on late developments.

an adjustable TVI filter for 300-ohm line

Bill, K5BDZ, has used an interesting TVI filter for a number of years with good success. It is the basic *handbook* filter which has been around for a number of years, except that Bill has changed the fixed capacitors to variable ones (**fig. 1**). The filter can be made up on a piece of perf-board for a trial run and later placed in a shielded box, if desired. The filter is installed at the 300-ohm TV receiver antenna terminals, or close to the tuner. Simply adjust the trimmer capacitors for maximum TVI rejection. If the filter is in a box, holes drilled in the cover will permit adjustment.

Bill has built about forty of these filters and has had good luck with all of them.

a construction project for beginners

Shown in **fig. 2** is a simple all-wave receiver for beginners that was designed by the U.S. Bureau of Standards. It can be easily built in a wooden cigar box by even the inexperienced. Designed about 1920, this venerable old circuit was featured recently in *Practical Wireless* (England). The magazine had this to say about the circuit:

"The object is to contact the crystal gently with the catswhisker in the hope of finding a sensitive spot which would cause rectification of the signal to take place. Many gentle applications were generally required before the spot could be found. Unfortunately, the slightest vibration would usually destroy this delicate setting, and the tedious business would have to be repeated."

And you think you have problems?

ham radio

10 GHz ultra stable oscillator

Frequency stability of 3.0 ppm is yours in a GaAs FET microwave source

The original work for the oscillator described in this article was for use as an LO in a DBS (direct broadcast satellite) downconverter. The frequency of oscillation is 10.76 GHz and all test data given is for this frequency. The 10.0 - 10.5 GHz Amateur band can also be easily accessed using this same oscillator. With correct selection of dielectric size, the same circuit can be made to oscillate from 7 to 16 GHz.

There are **two key elements** which make the application of a simple, stable microwave oscillator possible. The first is the GaAs FET (gallium arsenide field effect transistor); the second is a temperature-stable, high-dielectric, low-loss material (barium tetratitanate). The GaAs FET used for this oscillator is an ALF3000/ALF3003 and the dielectric material is type D8512 by Trans-Tech, Inc., Gaithersburg, Maryland.

The superior microwave performance of GaAs MESFETS is well documented, and usable gains are now possible from dc to over 40 GHz. The Alpha ALF3000 GaAs FET has over 9 dB of gain available at 10 GHz and simplifies the conditions for oscillation. It also has moderate power-handling capability and achieves + 17 dBm (50 milliwatts) of output power at 10 GHz with an efficciency of almost 28 percent.

Oscillator stability of better than 3 parts per million over a temperature range of -20 to $+60$ degrees C results from using the Trans-Tech dielectric. This type D8512 material has a Q_u (unloaded Q) of over *3000* at 10 GHz and is responsible for the high stability. **Fig. 1** shows a rough selection of material size for a given aspect ratio (diameter/height ratio) versus frequency. Final dielectric size will depend on what the housing (cavity) shape is and how it affects the resonant frequency.

Fig. 2 shows how the physical spacing of the dielectric from a cavity wall varies the frequency of oscillation. The oscillator has a large mechanical tuning range and therefore a stable housing (cavity) is a must for maximum stability.

A quick estimate of the frequency stability performance of any dielectric from barium tetratitanate can be obtained from knowledge of the dielectric's relative permittivity change with temperature. This constant, when divided by two (with reverse sign), is approximately equal to the frequency stability of the material impressed in a metal sandwich. A further improvement by a factor of two is achieved when 1/8 wave (length) of air spacing is included between the metal and the dielectric material.

There is a trade-off involved in Q_L, resonant frequency, and tuning range available in any given cavity size. Frequency resonance of the dielectric goes up as it is brought closer to a wall and Q_L goes down. Smaller cavity sizes also have the same effect.

Fig. 3. shows the quality of signal produced. Fm noise is extremely low and typically less than 0.1 $Hz/\sqrt{(Hz)}$ at 100 kHz off carrier.

It should be mentioned that many modes of oscillation are available, and care has been taken so as to couple only magnetically into the dominant TE01 mode.

In high dielectric material, the lines of magnetic flux are more tightly contained than in lower dielectric material. **Fig. 4** shows a simple view of this magnetic coupling of the dielectric material to microstrip.

The ability to frequency modulate this oscillator is shown by viewing the dc supply voltage versus frequency characteristics in **fig.** 5. Also shown is the output power versus dc supply voltage.

Figs. 6 and **7** show the actual circuit and PC artwork needed to reproduce this oscillator. PC board material is Duroid D5880, 31 mils thick, 1/2-ounce

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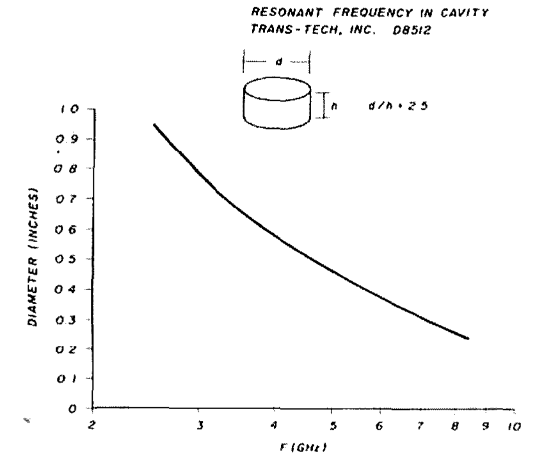

fig. 1. First order approximation for the diameter of a dielectric resonator versus frequency for a 2.5 aspect ratio, cavity with air space.

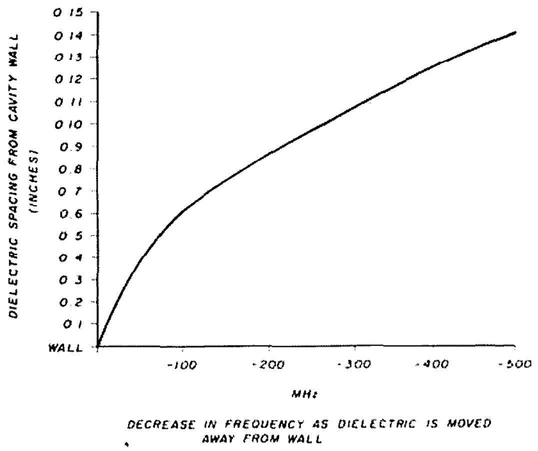

fig. 2. Dependence of mechanical air spacing versus frequency tuning.

copper on two sides with a dielectric constant of 2.55. This board was placed in an aluminum housing with conductive silver epoxy. A plated through-hole in the ground area at the end of the gate strip provides a good ground to the gate resistor. This resistor is a 50-ohm chip. If this is not in your parts box, a 1/8-watt carbon with short leads can also be made to work. GaAs FET devices used were in both chip and packaged form. In the packaged form (ALF3003) the gate lead was bent around the device to attach to the gate strip.

The design is very forgiving and has worked even with such problematic components at 10 GHz as 1/8-watt carbon resistors. The dielectric material was fixed in place on the PC board with Eastman 310 Superglue, which appeared to have no ill effects on performance once hardened. The circuit did not oscillate while the glue was wet and curing, however.

I believe that the circuit could be placed on Tef-Glas PC material and would perform as well. Fiberglass G-10 material could possibly be used for a housing, eliminating the aluminum machined housing.

conclusion

This puts 10 GHz well within reach of any Amateur with GaAs FETs and chip caps. (Even starting without any of these items, cost is still under $20.) The LO described here could be FM'ed on transmit with a reasonable power (50 mW) and used as the LO for a receiver with an i-f offset of, say, 28-30 MHz. The

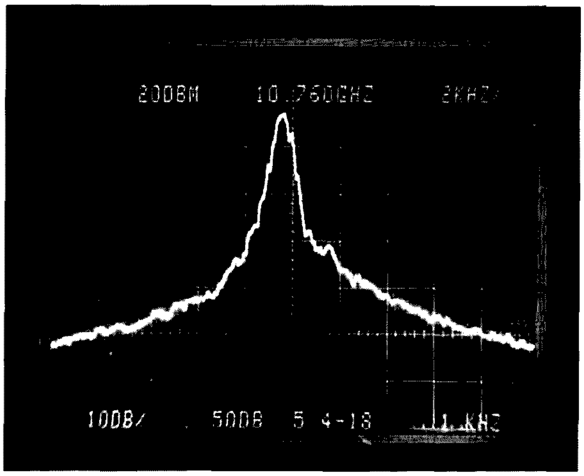

fig. 3. Spectral frequency response of a dielectric resonator oscillator.

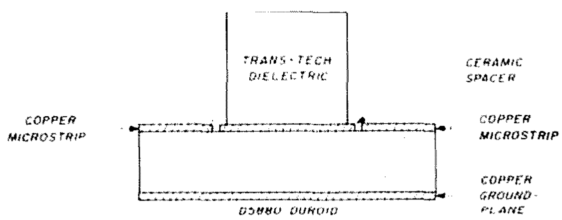

fig. 4. Fundamental mode coupling between the dielectric and microstrip line.

Note: While the ALF 3000/3003 is no longer commercially available, many other common FETs may be used in its place. —K8UR

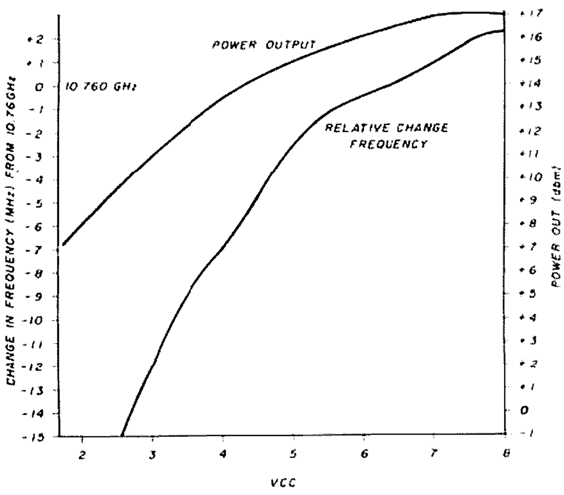

fig. 5. Dc tuning and power output versus supply voltage for the DRO.

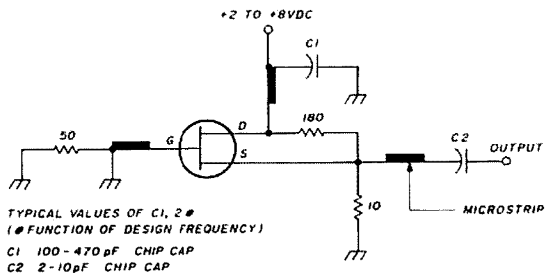

fig. 6. DRO schematic.

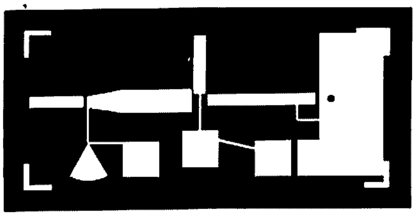

fig. 7. PC board artwork for the dielectric resonator oscillator (full scale).

cost is not much more than that of the FET alone, and it's my hope it will stimulate some 10-GHz interest.

ham radio

TRS 80 color computer for RTTY

Using the TRS 80 as a simple means of operating RTTY

Using the TRS 80 is one of the least expensive ways of getting on RTTY. A computer gives you flexibility in speed control and the possibility of split-screen operation, as well as programming capability. But whether to write the program or buy one is a question. I decided to buy.

After much looking, I finally located a program that did not cost too much yet had all the features I wanted. I loaded the tape into the computer and the TV monitor became alive with menus displaying Morse and RTTY commands. This was very nice, but all of my problems were not yet solved.

A TU and AFSK unit had to be built if the computer was going to be *able* to accept the RTTY *tones* emanating from the receiver. I remembered a TU design from the December, 1980, issue of *QST*. It was an excellent design, and modifying it seemed quite feasible.

Fig. 1 shows the schematic diagram as I modified it. I had no intention of running a TTY machine in series with the loop, so I deleted the optical isolator and the TTY driver circuit. The computer output line from pin 4 of connector P1 is an RS232-type of output (plus and minus swing around ground level). The RS232 level must be converted to a TTL level (zero to plus 4-volt level) fed to pin 9 of U2. U4 is an RS232-to-TTL converter; U1A in conjunction with U4 applies the correct polarity to the AFSK unit.

Pin 2 of P1 is the input line to the computer. **Data out** at pin 7 of U3 is connected to a 10-kilohm pullup resistor and the input of U1B.

A comment concerning the lack of audio filtering at the TU input: The Exar 2211 phase-locked loop is actually a filter. It works well without any bandpass filter at the input. But an audio bandpass filter with a bandwidth of approximately 400 Hz, with the MARK and SPACE tones centered in it, should improve TU

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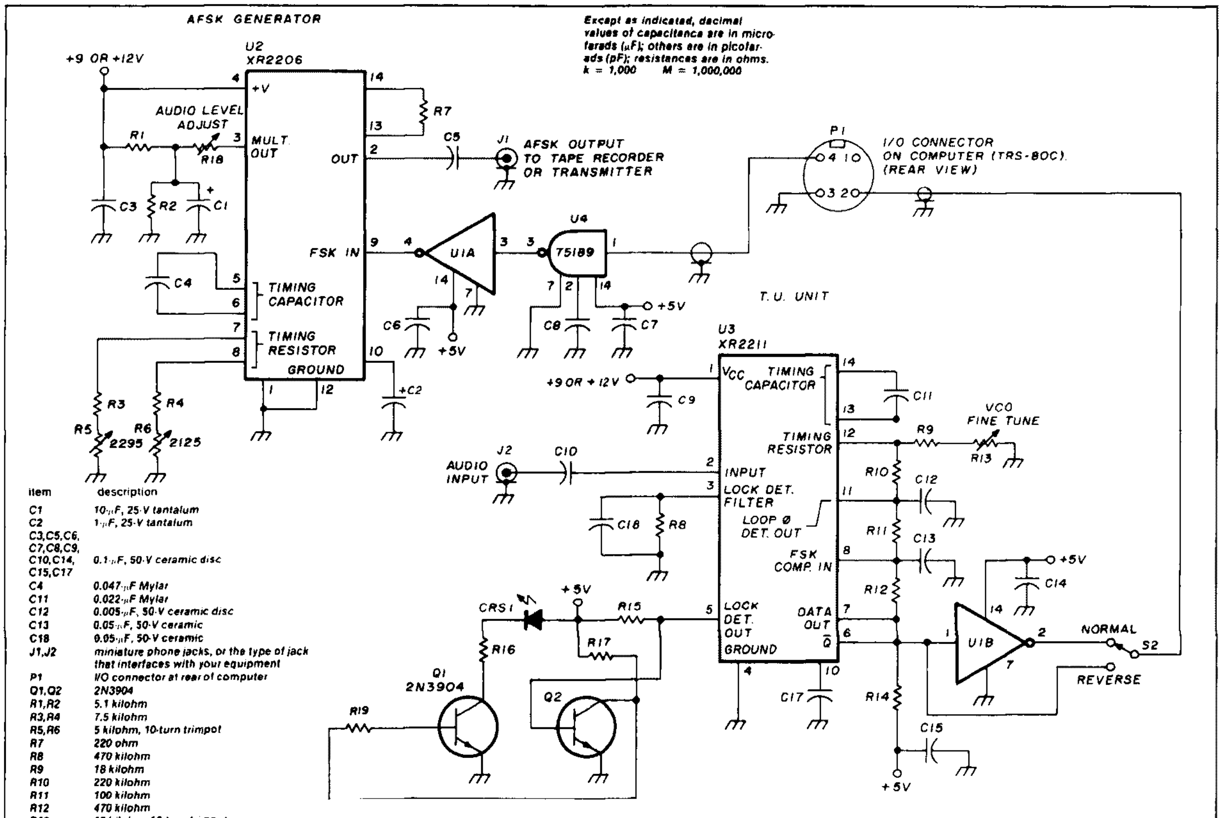


item	description
C1	10-μF, 25-V tantalum
C2	1-μF, 25-V tantalum
C3,C5,C6, C7,C8,C9, C10,C14, C15,C17	0.1-μF, 50-V ceramic disc
C4	0.047-μF Mylar
C11	0.022-μF Mylar
C12	0.005-μF, 50-V ceramic disc
C13	0.05-μF, 50-V ceramic
C18	0.05-μF, 50-V ceramic
J1,J2	miniature phone jacks, or the type of jack that interfaces with your equipment
P1	I/O connector at rear of computer
Q1,Q2	2N3904
R1,R2	5.1 kilohm
R3,R4	7.5 kilohm
R5,R6	5 kilohm, 10-turn trimpot
R7	220 ohm
R8	470 kilohm
R9	18 kilohm
R10	220 kilohm
R11	100 kilohm
R12	470 kilohm
R13	10 kilohm, 10-turn trimpot
R14,R15	10 kilohm
R16	470 ohm
R17	1 kilohm
R18	50 kilohm, single-turn trimpot
R19	3.3 kilohm
S1	SPST switch
S2	SPDT switch
U1	SN7404N
U2	EXAR 2206 (Exar Integrated Systems, Sunnyvale, California)
U3	EXAR 2201 (Exar Integrated Systems, Sunnyvale, California)
U4	SN75189N
CRS1	LED

fig. 1. Schematic of the TU as modified from design by WB2EWJ.

performance even more. My own receiver has a variable bandpass filter. I use the 400-Hz bandwidth and adjust the bandpass so that the tones are centered in the bandpass. This method is very effective with my older receiver, and I am sure it will be with modern transceivers as well.

To adjust the demodulator, disconnect pin 4 of U1A from pin 9 of U2. Ground pin 9 of U2 to simulate a MARK and connect J1 to J2. Slowly adjust the VCO fine tuning potentiometer until the LED lights; continue turning the potentiometer, counting the turns, until the LED goes out. Back the potentiometer off one-half the number of turns counted so that the VCO is set to the center of the lock range. This will get you sufficiently close to the optimum VCO setting, 2210 Hz.

Rf from the transmitter getting into the TU and video monitor can be a problem. I built my unit on a perforated copper-clad board enclosed in a shielded aluminum box. Be sure to bypass all power-supply inputs going to integrated circuits with 0.1-μF capacitors, as close to the power supply IC pin as possible. I/O lines should be miniature coax or any other type of shielded cable. The channel 3 modulator output from the computer should also be shielded. In conjunction with this, I found that a highpass TV filter was needed at the tuner input of the television set. In cases of extreme interference aluminum foil can be wrapped around the sides and top of the TV set; this shields the highpass filter and minimizes interference even further.

The end result was no discernible rf on the monitor screen on 80, 40, 20, and 10 meters. On 15 meters rf was noticeable but not objectionable; 21-MHz signal was probably getting into the i-f stage of the set.

The overall performance and reliability of this unit plus the computer is excellent. Signals very close to noise level can be copied with few errors.

reference

1. Michael J. DiJulio, WB2BWJ, "A State-of-the-Art Terminal Unit for RTTY," *QST*, December, 1980, page 20.

ham radio

DX FORECASTER

Garth Stonehocker, KØRYW

iast-minute forecast

The higher-frequency bands (6-30 meters) are expected to be best during the first and last weeks of the month, during periods of maximum solar flux. A slump may be experienced during the weeks in between, when the 10 and 15 meter bands are not able to support long skip, with only sporadic-E short-skip openings occurring. Try the lower frequency night bands: thunderstorm QRN isn't expected to be severe enough to wipe out evening and early-night DXing. Propagation disturbances are expected, however, around the 8th through the 14th. These geomagnetic-ionospheric disturbances (weak and fading signals) will probably not be intense but will probably last longer than usual (five to six days).

This year's June sunspot number is forecast to be only between 60 and 70. That is a significant drop from last year's value of 117. There will be quite a restricted frequency range for long-skip DX, even more restricted than usual during the summer. A total eclipse of the sun will occur in the Southern Hemisphere on June 11, from 0200 to 0715 UCT. The path begins near the Malagasy Republic, continues across southeast Asia, Indonesia, and Australia, to the west of New Zealand. Of note to moonbounce enthusiasts, a full moon will occur on the 25th and perigee on the 13th. And there will be a partial lunar eclipse on the 25th, which will be seen in the Americas and Australia. Summer solstice is June 21 at 2309 UT.

summertime DX

Each season of the year, the winter and summer solstices, the equinoxes, all produce their own distinct propagation characteristics, which can be put to work for DXing. Some of these characteristics can be turned to advantages, and some of the disadvantages can be circumvented. In the summer, the sun is more directly overhead in the Northern Hemisphere. The production of ions and electrons increases in the D, E (short skip) and lower F regions of the ionosphere. Greater concentrations of ions cause higher, lowest-usable-frequencies (LUFs), resulting in reduced signal strengths during the daylight hours. The higher number of E-region ions accumulate, producing sporadic-E (Es) layers; the F region, then lacking its ions (and electrons), provides the lowest maximum usable frequency (MUF) season of the year.

Summer also means more hours of daylight available for operating the higher DX bands. Summer thunderstorms are caused by air-mass heating because of the hot daytime temperatures. That makes these storms very different from storms caused by frontal passages, which occur during the equinoxes and in winter. Therefore QRN noise is now more common in the evenings and early night, after thunderstorms build up, and lasts until they dissipate.

Good DX on the lower bands can occur in the evening or after local midnight. Try short skip, since greater signal strength via Es might be effective in overcoming QRN. During the day, DX operation will be restricted to high-power transmitters capable of overcoming the high signal absorption of skywave and for good ground wave propagation. If you choose to use the longer daylight hours on the higher bands, despite the restricted operating frequency range for long skip (lower MUF and higher LUF), the evening is best, as the LUF falls with the sun and the MUF on 20 lasts into the night. If you'll settle for possible short-skip openings on 10 or 15 meters, try midday local time for the highest probability of openings.

band-by-band summary

Ten and fifteen meters should provide good daytime openings to the southeast, south, and southwest, using F-region long-skip hops of 2500 miles (4000 kilometers). Short-skip hops of 1200 miles (2000 kilometers) via sporadic-E should be possible during many days of the month in the above directions near local noon, and east and west before and after noon. Don't expect to find much one-hop trans-equatorial DX during disturbed periods this time of the year.

Twenty and thirty meters will be open to some parts of the world for nearly twenty-four hours a day. If 20 isn't useful some nights, 30 meters probably will be. Sporadic-E propagation will fill in the pre-sunrise dip in usable frequencies during many mornings to help make round-the-clock openings possible. The direction of the openings will be similar to those for 10 and 15 meters, plus the northern paths indicated on the chart.

Forty meters will provide the best DX conditions from sunset until just after sunrise, although static levels may be high at times. Watch for local storm passages and operate near sporadic-E peaks around sunrise and sunset (particularly at sunrise, when fewer thunderstorms are around).

JUNE

WESTERN USA

GMT	PDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	5:00	20	20	30	15	15	10	10	20
0100	6:00	20*	20	30	15	15	10	10	15
0200	7:00	20*	20	30	15	20*	10	10	20
0300	8:00	20	20	30	15	20	10	10	20
0400	9:00	20	20	40	20	20	10	10*	15
0500	10:00	20	20	40	20	20	10	15*	15
0600	11:00	20	20	20	20	20	15	15	15
0700	12:00	20	20	20	20	20	15	15	20
0800	1:00	—	20	20	20	20	15	15	20
0900	2:00	—	—	20	20	20	15	15	20
1000	3:00	—	—	—	—	20	15	20*	20
1100	4:00	—	—	—	—	—	20	20	20
1200	5:00	—	—	—	—	—	20	20	20
1300	6:00	—	—	—	20	—	20	—	—
1400	7:00	—	—	20	15	—	20	—	—
1500	8:00	—	20	20	15	—	20	—	—
1600	9:00	20	20	20	15	—	20	—	—
1700	10:00	20	20	20	15	—	15	20	—
1800	11:00	20	20	20*	15	20	15	20	20
1900	12:00	20	20	20*	15*	20	15	15	20
2000	1:00	20	20	20*	15*	15	15	15	20
2100	2:00	20	20	20	15	15	10	15*	20
2200	3:00	20	20	20	15	15	10	10	15
2300	4:00	20	20	30*	15	15	10	10	15

MID USA

GMT	MDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CDT
0000	6:00	15	20	30	15	15	10	10	15	7:00
0100	7:00	15	20	30	15	15	10	10	15	8:00
0200	8:00	15	20	30	15	20	10	10	15	9:00
0300	9:00	20	20	30	15	20	10	15	15	10:00
0400	10:00	20	20	40	20	20	15	15	20	11:00
0500	11:00	20	20	40	20	20	15	15	20	12:00
0600	12:00	—	20	20	20	20	15	15	20	1:00
0700	1:00	—	20	20	20	20	15	15	20	2:00
0800	2:00	—	20	20	20	20	15	15	20	3:00
0900	3:00	—	20	20	20	20	20*	20*	20	4:00
1000	4:00	—	—	20	20	20	20*	20	20	5:00
1100	5:00	—	—	—	20	—	20	20	—	6:00
1200	6:00	—	—	—	—	—	20	—	—	7:00
1300	7:00	—	—	—	20	—	—	—	—	8:00
1400	8:00	20	—	20	15	—	—	—	—	9:00
1500	9:00	20	20	20*	15	—	20	—	—	10:00
1600	10:00	20	20	20*	15	—	20	20	—	11:00
1700	11:00	20	20	20*	15	—	20	20	—	12:00
1800	12:00	20	20	15	15*	20	20	20	20	1:00
1900	1:00	20	20	20*	10	20	15	20	20	2:00
2000	2:00	—	20	20	10	15	15	15	20	3:00
2100	3:00	—	20*	20	15	15	15	15	20	4:00
2200	4:00	—	20*	20	15	15	15	15	20*	5:00
2300	5:00	—	20	20	15	15	15	15	15	6:00

EASTERN USA

GMT	EDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	8:00	15	20	30	15	30	10	10	15
0100	9:00	15	20	30	15	30	10	15*	15
0200	10:00	20*	20	30	15	30	15	15	15
0300	11:00	20	20	30	15	30	15	15	15
0400	12:00	20	20	40	20	30	15	15	20*
0500	1:00	20	20	20	20	20	15	15	20
0600	2:00	20	20	20	20	20	15	15	20
0700	3:00	—	20	20	20	20	15	20	20
0800	4:00	—	20	20	20	20	20	20	20
0900	5:00	—	20	20	20	20	20	20	—
1000	6:00	—	—	20	20	—	20	20	—
1100	7:00	—	—	—	—	—	20	—	—
1200	8:00	—	—	—	—	—	—	—	—
1300	9:00	—	—	—	15	—	—	—	—
1400	10:00	20	20	15	15	—	—	—	—
1500	11:00	20	20	15	15	—	20	—	—
1600	12:00	20	20*	15	15	—	20	20	—
1700	1:00	20	20	15	15*	—	20	20	—
1800	2:00	20	20	15	10	15	20	20	—
1900	3:00	—	20	15	10	15	15	20	20
2000	4:00	—	20*	15	10	15	15	15	20
2100	5:00	—	15	20	10	15	15	15	20
2200	6:00	20	15	20	15*	15	15	15	20*
2300	7:00	20	20*	20	15	20	10	15*	15

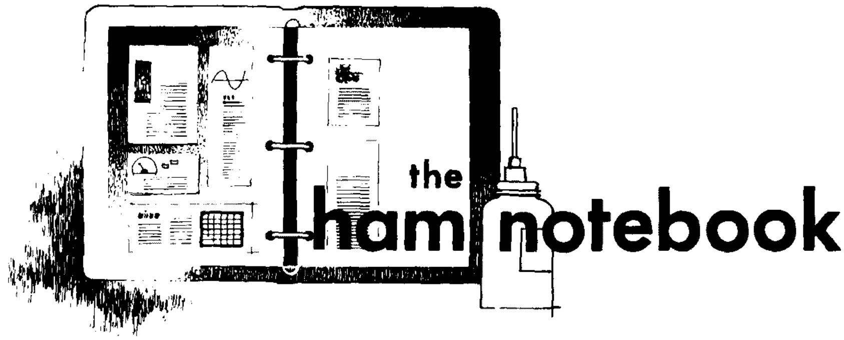
*Look at next higher band for possible openings.

Eighty meters during hours of darkness until sunrise can have DX openings to areas of interest. Static from thunderstorm activity, both long distance and local, may limit working the rare ones when propagation is otherwise all right. Coastal stations usually have more favorable propagation paths under summer conditions than do inland stations. Sporadic-E propagation around sunrise and sunset is good for this band also. Daytime work will be limited to within about 200 miles (360 kilometers).

One-sixty-meter DX activities really require a lot of work this time of year. During hours of darkness, between storm-front passages, you may work 1000 miles (1600 kilometers) if your ears hold up against the thunderstorm QRN. DXing in the predawn hours, after the thunderstorms have dissipated, may be the answer.

(Unfortunately an important path "looking" toward the west normally exists at this time; thunderstorm activity might not yet have subsided. — Editor)

ham radio



nomogram design of custom resistors

In Amateur work, it's sometimes necessary to fabricate a "customized" resistor in order to effect a quick design or repair. Though the right part may eventually be available, lengths of wire (in the right sizes, of course) may do in a pinch. The trick is to choose the right size wire and the right length (given the composition of the wire) to come up with the proper overall resistance. The equation itself is fairly simple:

$$R = \frac{\rho L}{1000K}$$

where R = resistance in ohms
ρ = resistivity of copper in ohms per 1000 feet at 20 degrees C (from wire tables)
L = length in feet
K = the ratio of the resistivity of copper to the resistivity of the wire's material

The nomogram simplifies solutions to the equation, in that resistivities have been converted to AWG (B&S) wire sizes. Also, the resistivities of various materials are given (**fig. 1**).

Two scales for resistivity and resistance are shown. If the outer resistivity scale is used, the outer resistance scale must also be used, and vice versa. The Length and Resistance scales may be changed proportionately. That is, if you need a range of the lengths of from zero to ten feet, the ranges of the resistances would also have to be reduced by a factor of ten — resulting in maximum resistances of 0.5 and 5.0. The relative resistivities of other materials may be plotted similarly. The chart also has usefulness in the design of control circuits, where the resistance of long runs of wire may be critical. Here's an example of how the chart is used: *Problem:* What length of No. 26 copper wire is needed to produce a resistance of 0.9 ohm? *Solution:* A line is drawn connecting copper on the Material scale to the proper wire size on the *outer* Resistivity scale. A second line is drawn from 0.9 on the *outer* Resistance scale (through the intersection of the first line and the diagonal), and it intercepts the Length scale at about 22 feet.

James McAlister, WA5EKA

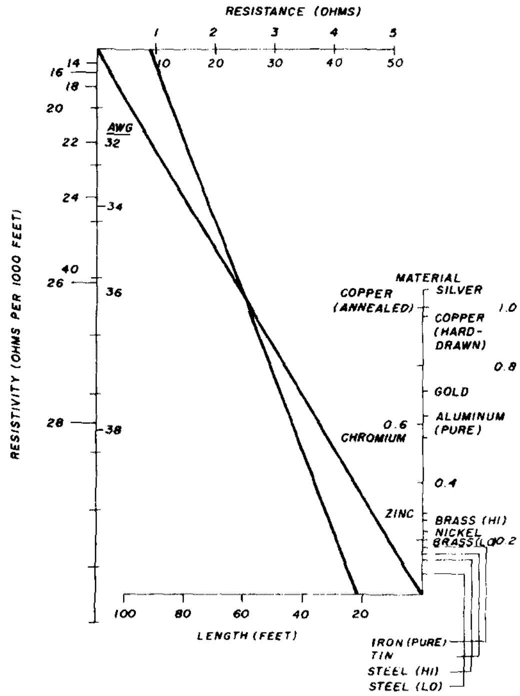


fig. 1. Nomograph for designing custom resistors.

operating the Triton IV on 30 meters

The TenTec Triton IV, which features QSK, can easily be converted for use on the new CW band, 30 meters. All that's needed is the addition of two switched capacitors and the modification of an outboard unit, the model 240 160-meter converter. Since many hams are more willing to work on an accessory than on a complex transceiver, this approach is very desirable.

Triton mods first

Remove the top cover and locate the bandpass filter board (No. 80291). Refer to the manual to find T2, the 14-MHz bandpass filter. C7 and C8 are within the T2 can and must be shunted by 56 pF of additional parallel

capacitance to lower the filter passband from 14 to 10 MHz. The terminals of C7 and C8 may be accessed without disturbing the can. Small wire-wrap wire connects C7 and C8 to a DPST N.O. (normally open) reed switch (taped to the top of T1, 2, 3, and 4). The 56-pF capacitors are soldered between the reed switch and the ground foil of the board. The reed switch is externally actuated by plac-

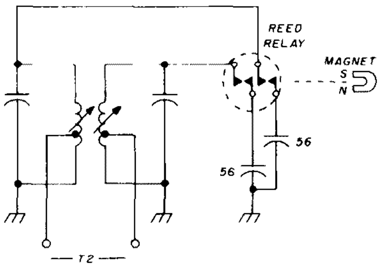


fig. 1. Magnet pulls in additional 56-pF capacitors, lowering the Triton IV bandpass filter to 10 MHz.

ing a magnet on top of the unit in a specific location; removing the magnet returns operation to 14 MHz. This is illustrated in **fig. 1**.

modifying the 160-meter converter

Relay K2 must be disabled on the converter filter board (No. 80328). Normally, K2 inserts a lowpass filter in the transceiver output line. This is necessary because the 3.5-MHz band lowpass in the Triton itself would allow second harmonic energy from 160-meter operation to pass along with the fundamental. This filter is not needed for 30-meter operation and is eliminated by opening the 12-volt switched line at pin 4 of the 80328 board. Don't be concerned about not having a filter in the line when 30 meters is being used. The internal Triton output filter already does that job.

Most of the work in modifying the model 240 is done on the mixer board No. 80327. Replace Y1 with a 4.0-MHz crystal. Inexpensive microprocessor crystals work well. Remove C18 and C21 from the circuit and replace C17 with approximately 3 to 4 pF. Remember that C17 now conducts 10-MHz rf, not 1.8-MHz. Also, we're already decreasing the Q of these tuned circuits by altering the L/C ratio. Consequently, only a small coupling capacitor is needed. Repeak C19 and C20 to receive signals between 10 and 10.5 MHz with the Triton bandswitch set for 20 meters: 10.0 corresponds to 14.0 on the dial.

IC1 on the mixer board is now fed by a 5.0-MHz VFO and a 4.0-MHz crystal oscillator. T1 needs to be centered at 1.0 MHz, the difference frequency. Remove T1 and C5, C6, and C7. T1 should be replaced with a 455-kHz i-f can from a tube-type radio, that is, a transformer with two high-impedance windings.

Before installing the new i-f can, open it and determine the value of capacitance. Remove this capacitor. The externally added capacitors will have approximately half the old value and resonate both windings of the transformer at 1 MHz. Some cans have a low-impedance tap on one winding or the other. The tapped winding becomes the output winding and the **MIX OUT** pin is fed directly

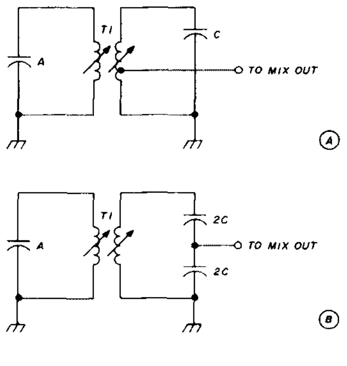


fig. 2. Two methods of obtaining a mixer output signal from the new 1-MHz i-f. (A) low impedance tap; (B) capacitive voltage divider.

from this tap. If your new can has no tap, just make a voltage divider to feed **MIX OUT**: simply obtain two capacitors of twice the value needed to resonate the output winding; *series* connect these two capacitors and tap at the junction for the **MIX OUT**. See **fig. 2**.

The desired mixer output voltage level is set as high as possible without driving the buffer transistor stage (80329) into distortion. Observe the output waveform of the buffer while peaking the primary and secondary of transformer T1.

For normal operation of the Triton, leave the Model 240 panel switch in the 3.5-30 position. For 30-meter operation, turn the Model 240 switch on and put the magnet in place on the Triton top cover (or switch in the extra capacitors via whatever switching mechanism you've chosen). For increased sensitivity, transformers T2 and T3 can be rewound for narrower bandwidth.

Raymond Henry, Jr., AA4LL

using the Astro 103 as a frequency counter

The Astro 103 transceiver's digital readout may be used as a general frequency counter without modifying the equipment. A phono jack labeled **EXT LO** on the rear panel is for connecting an external local oscillator, to obtain general coverage between 1.5 and 30 MHz. The **EXT LO** input circuitry is shown in **fig. 1**.

Because counter sensitivity is less than 10 millivolts, precautions must be taken to limit damaging voltages which might be applied accidentally. This is accomplished by assembling a simple, small, coaxial-line test lead to a phone plug, with a 10-pF, 1000-volt disc capacitor in series with the center conductor. This small value also prevents the overloading of any sensitive circuits *being measured*.

To measure an external frequency,

sending CW: a digital approach

A simple way to clean up your CW sending technique

International Morse code appears to be fairly simple on the surface. Many licensed Amateurs can testify to having mastered it with varying degrees of proficiency. When compared to modern digital codes, Morse code is really quite complex: two different on elements, dots and dashes, are used along with three different off elements to represent many letters, numbers, and special characters and symbols. Transmission speeds vary widely in a single transmission and dots, dashes, and spaces often become mashed in a stream of indecipherable information. This article details the use of digital circuits to accurately time and filter out-going Morse code sent with a straight key or other mechanical key so that dots and dashes are reproducible and well-timed. No new hand motions are required.

Someone who has listened to the CW ham bands for a short time will realize there are Amateurs who send code poorly. You may be unpleasantly surprised by your own CW-sending skills if you record one of your contacts and replay it. The main problem in sending good code is shown in **fig. 1**. A perfect transmission is shown at the top, and a typical on-the-air transmission is shown below. Note that real dots and dashes vary in length and the spacing is less than perfect. While the human brain is capable of adjusting for most of these changes, it is difficult to decipher a transmission in which dots and dashes are almost the same length and spaced in a semi-random fashion. Electronic paddle keyers, keyboards and computers are one answer, but they require mastery of new hand motions or skills and are frequently expensive. Many Amateurs would like to continue using their mechanical keys and improve their code-sending skills.

defining the circuit

The problem is to design a circuit that will accurately time dots, dashes, and spaces so they sound perfect on the air. Before the circuit can be designed, we need to specify what it is supposed to do. First we define the dot:dash:space ratio as being 1:3:1, the standard generally accepted for good Morse code transmissions. While many hams use weighting

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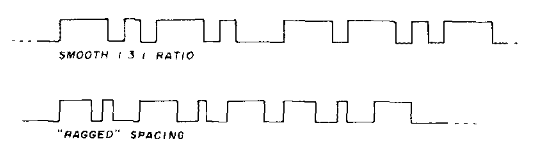


fig. 1. Perfect code versus real hand-sent code for CQ.

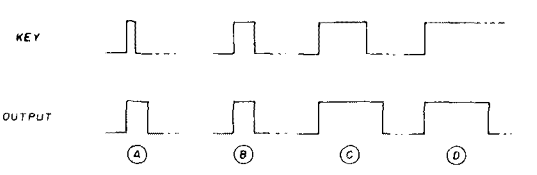


fig. 2. Dot and dash timing operations for the keyer circuit.

to vary the dot:dash ratio, it is better to use a well-known standard. Next, the dot/dash operation must be defined (shown in **fig. 2**). There are four actions:

A. If the key is closed for a very short time, a self-completing dot is generated.

B. If the key is closed and held closed up until the end of a dot, the same length dot is generated.

C. If the key is closed and held closed beyond the dot length, a self-completing dash is generated.

D. If the key is held closed for a long time, only a single dash is generated.

After each dot or dash, an off time of one dot-period is enforced to prevent code elements from being produced too tightly. If the key is activated during the off time, the action is remembered and acted upon as soon as the off time is finished. Perfectly-timed and spaced code can be generated by slightly leading the actions of the circuit.

using monostables

A simple timing circuit can be built using monostables, as shown in **fig. 3**. Monostable 1 is triggered by the key closure and generates a pulse one unit long. The negative-going edge of this dot pulse will trigger monostable 2 if the key is still pressed at this time. Monostable 2 generates a pulse that is two units long; when added to the length of the dot pulse, a dash pulse of three units is generated. The outputs of monostables 1 and 2 are gated to generate the Morse code output.

The output of monostable 2 is also used to disable (turn off) the key input to monostable 1 so it cannot be re-triggered during a dash. A third monostable is preset for one dot-time to generate the minimum required off period between dots and dashes. Monostable 1 is re-enabled so that the process may be repeated at the end of a dot or dash. While this circuit is useful in explaining the operations we would like, it does have some limitations: three monostables must be adjusted to change the speed of transmission, and key closures during the off period are not recognized. The circuit can be made to work but is impractical.

using a sequencer

A more reasonable approach uses a master clock

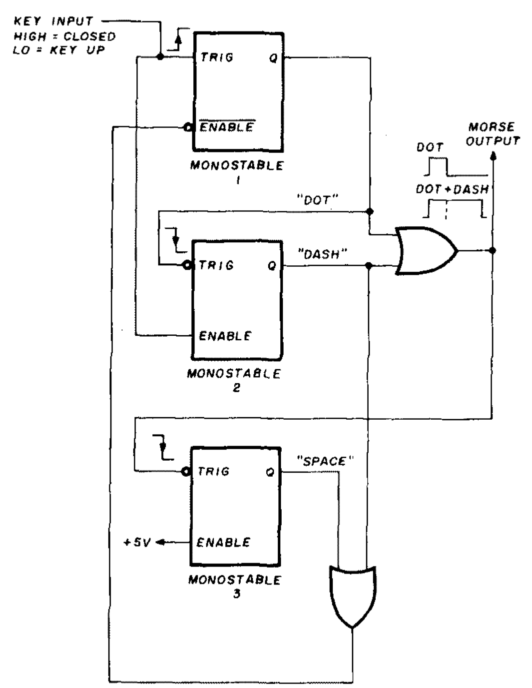


fig. 3. Using three monostables to filter hand-sent code.

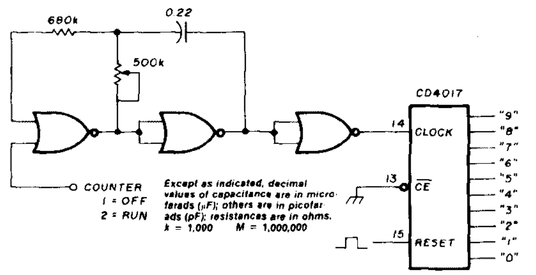


fig. 4. A counter/divider and gated clock used as a sequencer.

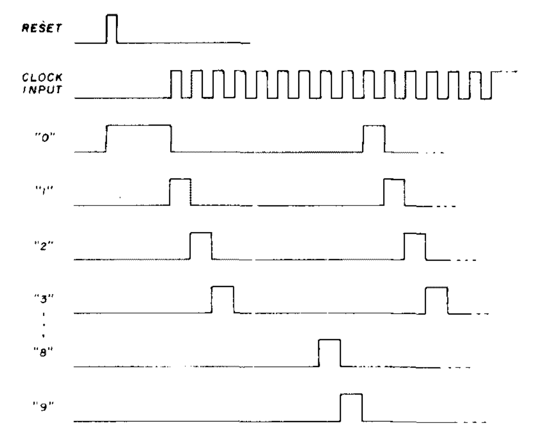


fig. 5. Timing diagram for a 10-stage sequencer circuit.

to sequence through a series of code-generating and condition-testing steps. Only the frequency of the clock need be adjusted to vary the code timing. The circuit described here was designed using digital complementary metal-oxide semiconductor (CMOS) integrated circuits to reduce power consumption.

As shown in **fig. 4**, the heart of the circuit is a CMOS CD4017 decimal counter/divider integrated circuit and a gated clock used as a sequencer. The CD4017 accepts input pulses at pin 14 and increments an internal count by one for each pulse. Only one of the ten outputs can be a logic one, indicating the state of the counter. Thus, ten external circuits can be turned on and off in a regular sequence, governed by the frequency of the clock signal applied to the CD4017. This is illustrated by the timing diagram in **fig. 5**. This sequencing circuit is used to generate the dot/dash/space times and to sequence various circuit elements that can test the key input for changing on and off conditions.

If the output of a free-running square-wave oscillator is controlled with a gate as shown in **fig. 6**, the first cycle output will be of arbitrary length. All subsequent cycles will be of equal length. This type of gated waveform cannot be used for accurate timing. The trick is to use an oscillator that is triggered or started by the enabling signal. The first clock cycle generated by a gated clock is always the same length as following cycles. When such a gated clock is used with the CD4017 sequencer circuit, the first sequence will be the same length as those that follow. The gated clock is simply a monostable that retriggers itself when gated on.

The Morse-generator portion of the circuit can be built from the sequencer and several CD4025 three-input NOR gates, shown in **fig. 7**. The dash input control line determines whether or not a dash is sent. If $DASH = 1$, then a dot is sent. If $DASH = 0$, the sequence is extended and the dot is stretched into a dash. A dot is always sent on sequence start.

generating a dot

The rest state of the sequencer is a logic one at the 0 output with logic zeros at all other outputs. This is the sequencer state when the gated clock is enabled. The first edge of the clock signal increments the internal sequencer count, moving the logic one to the 1 output. The logic one at the 1 output causes the lower NOR gate to output a logic zero. The sequencer moves the logic one to the 2 output on the next clock cycle and the output of the lower NOR gate goes back to the logic one state. This generates a logic-zero dot one clock cycle long. There is no other effect: any action through the upper NOR gates is blocked by the logic one on the dash control line. The time of the sequencer 2 output generates the off period (one clock cycle) at the end of each dot. Output 3 is the dot reset line (described in detail shortly). It resets the dot-generating sequence.

generating a dash

Let's see what happens when a logic zero on the dash control line allows the sequencer outputs to pass through the two upper NOR gates. The se-

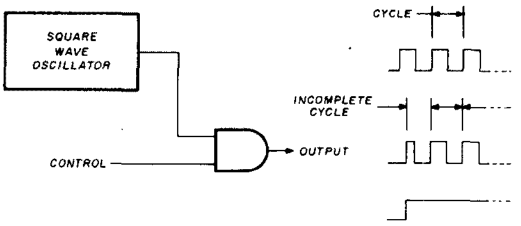


fig. 6. A simple gated free-running oscillator. Note that the first gated cycle may be incomplete.

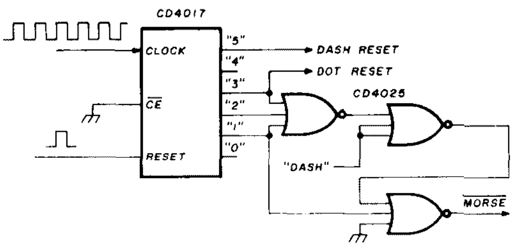


fig. 7. Using the sequencer circuit to generate perfectly timed dots and dashes.

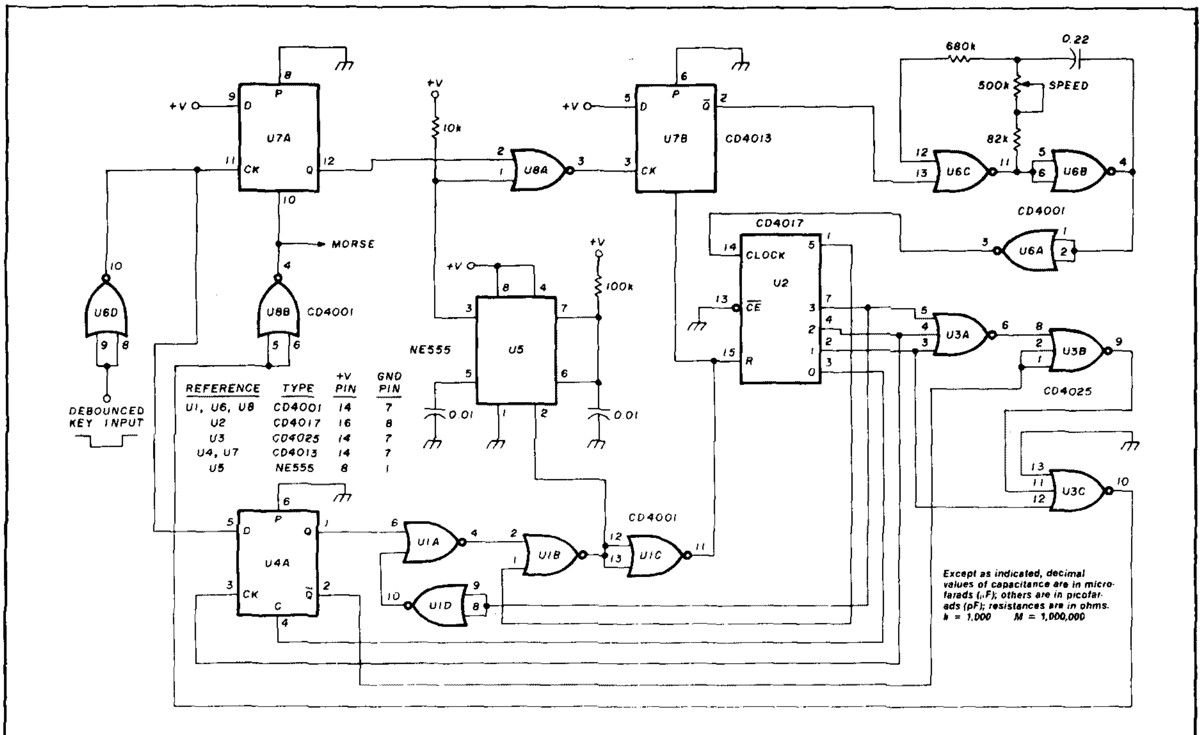


fig. 8. The complete keyer circuit.

quencer has been reset to 0 and the first clock pulse moves it to the 1 state. The output of the lower NOR gate goes to a logic zero once again. Since the two upper NOR gates are now enabled by the dash control line, the logic one outputs at 2 and 3 are passed through the lower NOR gate to generate a logic zero that is present for three clock cycles. This forms a dash exactly three times as long as a dot. Output 4 is not used and this sequence time provides the off period at the end of each dash. Output 5 is the dash reset line which resets the dash-generating sequence.

completing the circuit

Generating the dot and dash sequences is straightforward. Other circuits are used to detect the key closure and decide whether to send a dot or to extend the sequence and send a dash. A series of CD4013 D-type flip-flops and CD4001 two-input NOR gates are used for this, and shown in **fig. 8**.

The key input clocks a CD4013 flip-flop, U7A. The $\overline{Q}$ output from this flip-flop goes through a NOR gate to clock flip-flop U7B. U7B-2 turns on the gated clock to start the timing sequence. The Morse output from the sequencer/gating circuit is inverted and applied to the reset input of flip-flop U7A. This clears U7A and holds it in the cleared state for the length of any Morse element (dot or dash) being generated. The circuit is insensitive to any keying actions while it is generating a dot or dash. Since the clearing signal is the dot or dash, it is not present during the enforced off period and the circuit can detect another key closure during this time.

Several of the sequencer outputs are used to control another flip-flop, U4A, and four two-input NOR gates, U1. This part of the circuit determines whether or not to change from a dot to a dash sequence. Here is how it works: flip-flop U4A is used to determine whether or not the key is still pressed at the end of the dot now being generated. The key input provides the data (D input) signal to the flip-flop clocked at the start of the off period following a dot. If the key is open (as it would be for a properly sent dot) the state of the flip-flop will not be changed; the Q output remains a logic zero and the $\overline{Q}$ a logic one. So, when sequencer output 3 becomes a logic one, it will be passed through the four NOR gates (U1) and reset both the sequencer and flip-flop U7B. This ends the dot-generating sequence. Remember that the 0 output is a logic one when the sequencer is reset.

The key is held closed to send a dash. When the positive edge of sequencer output 2 appears, it clocks the key-closed condition into flip-flop U4A and the sequence is modified. The $\overline{Q}$ output (dash) from the flip-flop enables NOR gate U3B so the dash

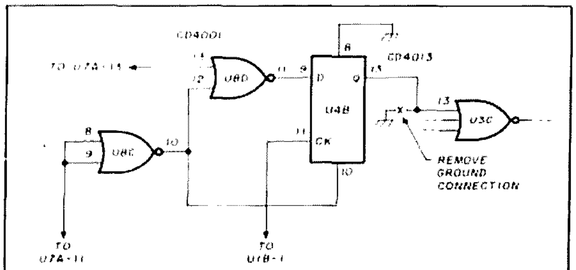


fig. 9. These circuits are added for the automatic tune-up feature.

sequence can be completed. The reset pulse normally generated by sequencer output 3 is blocked since the Q output from flip-flop U4A is now a logic one. This means that when the sequencer completes step 3, it will continue on to step 4 and then to step 5. The 5 output is now fed into NOR gates U1B and U1C to reset the system. When the sequencer returns to its reset condition, the logic one from the 0 output resets flip-flop U4A.

The same pulse that resets the sequencer counter/divider also resets flip-flop U7B, which controls the gated clock, turning the clock off. Since key-sensing flip-flop U7A may have detected a key closure during the enforced off time between elements, its condition must be tested and passed through to flip-flop U7B after the system has been reset. The sequence must be started again if a key closure is waiting. The testing operation is done by the NE555 monostable, U5, which is triggered by the reset pulse that clears the system. The pulse generated by monostable U5 is longer than the reset pulse. This allows the circuit to be completely reset before any new key-closure information is passed through to flip-flop U7B to restart the Morse-generating sequences. As mentioned previously, this allows the circuit to detect a key closure taking place during the enforced off time between dot and dash elements.

adding a tune-up circuit

Two unused gates and an unused flip-flop exist in the original circuit. These have been used to form an automatic tune-up circuit. Many hams like to make quick on-the-air adjustments to their transmitter or antenna, using key-down tuning for this. Most electronic keyer circuits use another switch or control function to constantly key the rig. If the keyer described here is used, you could only generate dots and dashes. The simple addition shown in **fig. 9** allows for constant keying.

This circuit checks to see if you still have your key closed at the end of a complete dash-generating cycle. If you depress your key and hold it closed, the keyer will generate a dash, a space and then go into a constantly-keyed mode so you can tune your rig. Releasing the key resets this operation so you can send code normally. No added tune control is needed when this tune-up circuit is used. Note: keep on-the-air tune-ups as short as possible! A complete keyer unit is shown in **fig. 10.**

The circuit described here follows your key operations instantaneously. There is no annoying dead-time or delay between your key closure and start of keying. The decision as to whether or not to send a dash is made on the fly. If you try and send too fast for the speed setting, you will immediately hear the result from your side-tone oscillator and can adjust your speed accordingly.

This circuit will generate accurately timed and spaced code for you and no new hand motions are required. Most hams take about 10 to 15 minutes to become accustomed to sending accurately-timed code with this keyer circuit. Since most of us are a bit inconsistent in our sending, the circuit will clean up the ragged edges of our code so it sounds almost perfect. Of course, it's up to you to generate the required spaces between characters and words.

learning the code

One of the reasons for designing this code-timing circuit was for hams to learn the sound of well-sent code and to learn sending good code with a straight key. Since you can only send code elements in the ratio of 1:3 with this keyer circuit, you quickly learn from the aural feedback whether you are sending good code or not. The ratio of 1:3:1 for dots, dashes and spaces is a bit difficult to master and this circuit can be used to great advantage in teaching newcomers the proper way to send code.

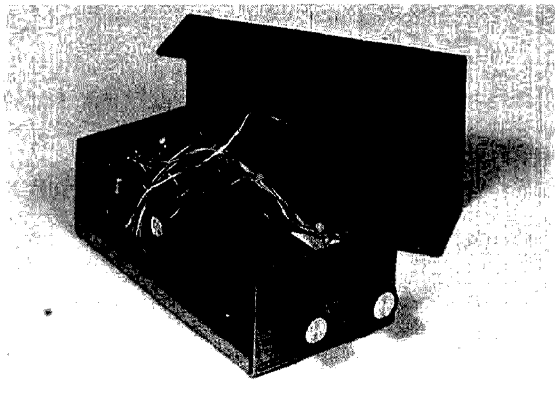


fig. 10. The completed keyer.

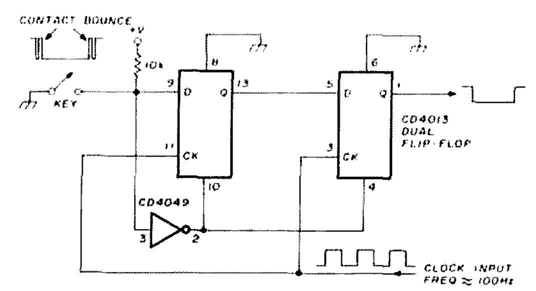


fig. 11. A sample switch debouncing circuit for single-pole single-throw switches.

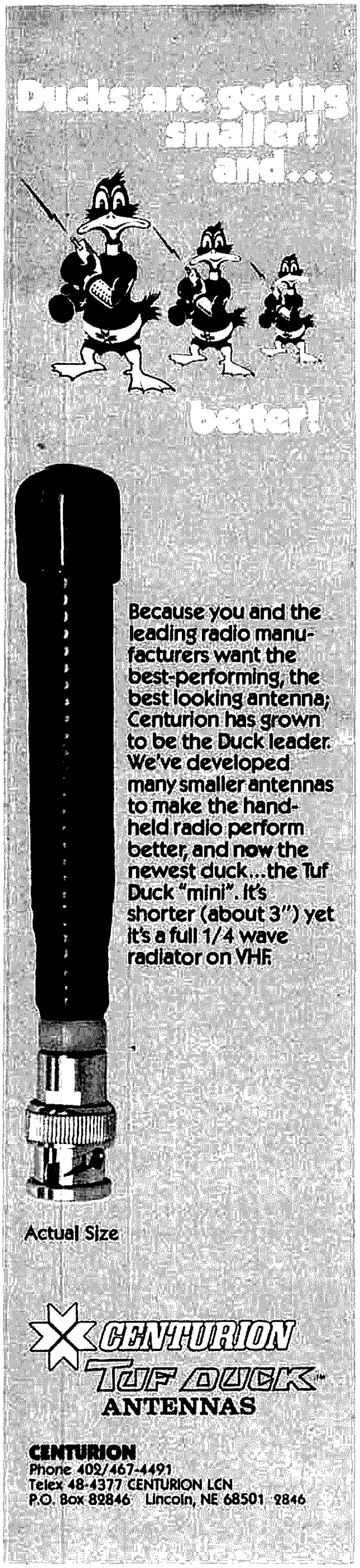
finishing touches

The circuit in this article needs a key-debouncer as well as a transmitter keying circuit. A sidetone oscillator can be added if desired. The debouncer circuit can be a simple RC network or clocked circuit to filter multiple-contact closures characteristic of mechanical switches. A typical debounce circuit is shown in **fig.** 11. The external clock is set at about 100 Hz for 100 ms bounces of the key. The transmitter-keying circuit will depend on your rig; a small reed relay with appropriate contact rating and transistor oscillator can be used. A sidetone oscillator is easy to build with a 555 timer. This is recommended for off-the-air use, particularly if you're helping someone get started in ham radio.

CMOS devices in this design allow a power supply that provides 5 to 15 volts. I recommend using at least 9 volts. Most modern solid-state rigs use 12 to 14 Vdc, quite adequate. I don't recommend plug-in battery eliminators as a power source unless voltage regulating and filtering circuits are added.

Power consumption is low but will increase if you decide to add a sidetone oscillator driving a small speaker. When the automatic tune-up circuit is used with the keyer, only a speed control is needed. The upper sending speed may be increased by shorting the fixed series resistor with the Speed control in **fig. 8**. The values given provide speeds in the range of 3 to 30 words per minute. A sidetone oscillator volume control may be added, plus a power switch. CMOS circuits are fairly immune to electrical noise but a metal enclosure is recommended to protect the circuit from RFI.

ham radio



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WANTED, MILITARY SURPLUS RADIOS. We need Collins 618T, 671U, ARC-94, ARC-102, MRC-95, RT-804A/APN-171, RT-712/ARC-105, ARC-114, ARC-115, RT-823/ARC-131 or FM622, RT-857/ARC-134, or Wilcox 807A, RT-859/APX-72, ARC-164, PRC-25, PRC-64A, PRC-66, PRC-68, PRC-70, PRC-74, PRC-77, PRC-104, RT-524/VRC, antenna couplers 490T, CU-1658A, CU-1669A, 490B-1, 690D-1. Top dollar paid or trade for new Amateur gear. Write or phone Bill Slep, 704-524-7519, Slep Electronics Company, Hwy. 441, Otto, N.C. 28763.

WANTED: Schematics-Rider, Sams or other early publications. Scaramella, P.O. Box 1, Woonsocket, RI 02895-0001.

TEKTRONIX plug-in units. EXC: 1A4, $175, 1A1, $125, others too. H-P meters, $15, spectrum analyzers, tube tester, oscillators, much more on huge spring-cleaning list. S.A.S.E. please. Joseph Cohen, 200 Woodside, Winthrop, MA 02152.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

ELECTRONICS BOOKS, engineering magazines, reference & databooks, radio magazines, many 25¢, some from 1930's. Also Test Equipment manuals and catalogs, machinist magazines, porcelain insulators, scope viewing hoods, much more on huge list. S.A.S.E. please. Joseph Cohen, 200 Woodside, Winthrop, MA 02152.

RUBBER STAMPS: 3 lines $3.25 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

RADIO WEST! High quality coverage communications receivers with Collins filters and other DX modifications! Catalog 50¢ (refundable). Radio West, 3417 Purer Rd., Dept HR, Escondido, CA 92025 (619) 741-2891.

WANTED: New or used MS and coaxial connectors, synchros, tubes, components, military surplus equipment. Bill Williams, PO #7057, Norfolk, VA 23509.

SELL: Ten-Tec Century 21 40W CW, 80-10 meter xcvr., mint $175.00. Yaesu FT202R, 2 meter HT with charger, very good $75. HAL CWR6700 RTTY reader with green video monitor, mint $525. WA2RCC, Victor Miller, 55 Knolls Crescent, Bronx, NY (212) 548-8077.

VERY in-ter-est-ing! Next 5 issues $2. Ham Trader "Yellow Sheets". POB356, Wheaton, IL 60189.

SELL: Yaesu FT 102 with narrow SSB filter. Matching speaker & MD-1 desk mike. Mint condition. $785.00. W6OWD (415) 728-7136.

CB TO 10 METER PROFESSIONALS: Your rig or buy ours — AM/FM/SSB/CW. Certified Communications, 4138 So. Ferris, Fremont, Michigan 49412; (616) 924-4561.

SELL COLLECTION of 30 years. New boxed receiving tubes only one dollar each. 4-1000 A Pulls only $50.00 each. Sell Classic Drake 1A triple conversion receiver $125.00. Collins 75A4 with 3 filters recently aligned $295.00. W5QJT, PO Box 13151 Coronado Station, El Paso, TX 79913.

HAMS FOR CHRIST — Reach other Hams with a Gospel Tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

SIGNAL GENERATOR 80 KC — 60 MC $35, HP200B Audio Osc. $30, Hallicrafters HT-18 $35, Hickock M/C tube tester $35, RCA WV-98B VTVM $35, B&K 440 CRT tester $20. K6KZT, 2255 Alexander, Los Osos, CA 93402.

PLANS, CIRCUIT BOARDS, AND KIT PARTS (author approved) for Leach's construction projects. Power amps, preamps, pre-amps and loudspeakers. Send SASE for information. Custom Components, Box 33193, Decatur, GA 30033.

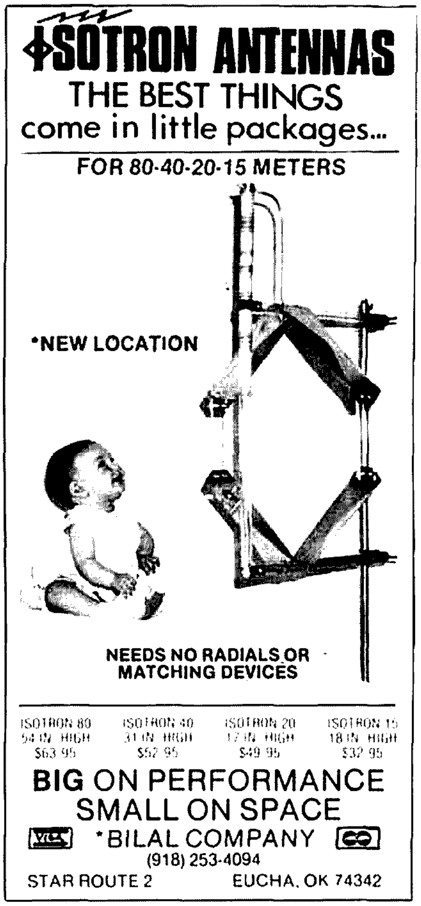
WANTED: Pro-pitch motor. State condition, best price. Jim, K3VJH, 2219 Ross Court, Silver Spring, MD 20910. (301) 565-2219.

FOR SALE— Complete HF station. Kenwood TS520 w/SP520 speaker. Hy-Gain M# TH3Mk2 Tribander. CDE M# AR-22R rotor and control. Rohn M# 25 tower (50'). All for $600 or will sell separately. Call Herb (WB3CVA) (717) 394-5723.

WANTED: Manual and circuit prints for TV cameras Fairchild TC-177RL and Concord MTC-21. WB0PXQ, Rt. 1, Brady, Nebraska 69123.

FOR SALE: Hercules 444, 1 KW amplifier. All solid state by Ten-Tec. Absolutely mint cond. $850.00. KA6EGX. (916) 694-2374.

IBM-PC RTTY & ASCII. SASE for full details. E. Altine, NE5S, 773 Rosa, Metairie, LA 70005.



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Coming Events

ACTIVITIES

"Places to go..."

CALIFORNIA: The Satellite ARC's Amateur Radio Swapfest and BBQ. June 19, 10 AM to 4 PM, Union Oil Company New Love Picnic Grounds south of Santa Maria off U.S. 101. Free admission to Swapfest. BBQ tickets — adults $7.50, children 6-12 $3.50, children under 6 free. Swap tables $2.50. Talk in on 146.34/94. For tickets and information: Santa Maria Swapfest, PO Box 2616, Orcutt, CA 93455.

COLORADO: SUPERFEST V sponsored by the Northern Colorado ARC, June 4, 8:30 - 3:30 PM, Larimer County Fairgrounds, Loveland. Family activities, technical talks and a swapfest. For information: Rick Hubbard, WA0DDC.

IDAHO: The Treasure Valley Amateur Radio Association's Hamfest, June 18, Payette. Games, swap tables, family fun and a banquet. Pre-registration $5.00. For more information SASE to PO Box 790, Payette, ID 83661.

INDIANA: The Lake County Amateur Radio Club's 11th annual "DADS DAY" Hamfest, June 19, Lake County Fairgrounds, Industrial Arts Building, Crown Point, 8 AM til ??. Tickets $2.50. Talk in on 147.84/24 or 52. For further information: Denny Tokarz, KA9FCG, 6930 Lindbergh, Hammond, IN 46323.

INDIANA: The Indiana State Amateur Radio Convention in conjunction with the Indianapolis Hamfest and Computer Show, Sunday, July 10, Marion County Fairgrounds, intersection of I-74 and I-465. Inside/outside flea markets. Separate computer show and flea market. Commercial vendors. Setup after 12 noon Saturday, July 9. Camper hookups available on grounds. Nearby motels. Gate ticket $4.00 entitles you to all activities. For further information: Indianapolis Hamfest, Box 11086, Indianapolis, IN 46201.

MARYLAND: The Frederick Amateur Radio Club's 6th annual Hamfest, June 19, 8 AM to 4 PM, Frederick Fairgrounds. Admission $3.00. YLs and children free. Tailgaters $2.00 extra. Gates open for exhibitors 8 PM, June 18. Overnight parking welcomed. Exhibitor tables $10.00 for the first, each extra $5.00. For more information: V.A. Simmons, KA3CVD, 7301 Pin Oak Drive, Middletown, MD 21769. (301) 371-5735.

NEW JERSEY: The Jersey Shore Chaverim are sponsoring the Jersey Shore Hamfest and electronic flea market, June 12, 9 AM to 3:30 PM, Jewish Community Center, 100 Grand Avenue, Deal. Admission $3 per person. Children under 12 and XYLs free. Refreshments available. Table $5. Tailgating $2.50. Reserve spaces by SASE and advance payment to Jersey Shore Hamfest, PO Box 192, West Long Branch, NJ 07764 by May 15. Talk in on 147.045 + .6; 146.52 simplex.

NEW JERSEY: The Raritan Valley Radio Club's 12th annual Hamfest, Saturday, June 18. Gates open 8:30 AM at Columbia Park, Dunellen. Sellers $3.00 each spot, no tables provided. Lookers $2.00 donation. Advance tickets purchased from any club member. Talk in on Club repeater, W2QW/R 146.025/.625 and 146.52 simplex. For further information: Bob, KB2EF or Mary, WA2JWS (201) 369-7038 — 10 AM to 10 PM.

NEW YORK CITY: The Hall of Science Amateur Radio Club's annual indoor/outdoor, rain or shine, Electronic Hamfest, Sunday, June 12, 9 AM to 4 PM, Municipal Parking Lot, 80-25 126th Street, (1 block off Queens Blvd.) Kew Gardens, Queens. Sellers donation $3.00; buyers $2.00; XYLs and kids free. Walk/talk in on 146.520. For information: Tony Russo, WB2OLB (212) 441-6545. John Powers, KA2AHJ (212) 847-8007.

NEW YORK: The Skyline Amateur Radio Club (SARC) Hamfest, June 18, 9 AM to 5 PM, Cortland County Fairgrounds, I-81, Cortland. Rain or shine, indoor/outdoor flea markets. Talk in on 52 simplex. Welcome all!

NEW YORK: The Genesee Radio Amateurs' third annual Batavia Hamfest, Sunday, July 10, 6 AM to 5 PM, Alexander Firemen's Grounds, Rt. 98, Alexander. Registration $2 advance, $3 at the gate. Large exhibit area, contests, OM/YL programs, food, boat anchor auction, overnight camping. Fun for all! Talk in to W2RCX on 6.52 or 4.71/5.31. For advance tickets (check payable to Batavia Hamfest) or information: GRAM, PO Box 572, Batavia, NY 14020.

OHIO: The Champaign Logan Amateur Radio Club's annual Hamfest and Flea Market, Sunday, June 12, Logan County Fairgrounds, Bellefontaine. Gates open 7 AM EDST. Tickets $1.50 advance; $2.00 at door. Tables $3.00 advance. For information, tickets or tables: Michael DeVault, KU8I, 7157 Road 158, East Liberty, OHIO 43319.

OHIO: The Goodyear Amateur Radio Club's 16th annual Hamfest, Sunday, June 12, Goodyear Wingfoot Lake Park near Akron. Picnic and flea market 10 AM to 5 PM. Family admission $2.50 advance, $3.00 gate. Outdoor flea market $2.00 per space. Dealers inside $5.00 per table (advance reservations suggested). For tickets and information SASE to: Don Rogers, WA8SXJ, 161 S. Hawkins Avenue, Akron, Ohio 44313. (216) 864-3665.

OHIO: The Battelle Amateur Radio Club (W8CQK) is sponsoring the 3rd annual Columbus Hamfest, Saturday, June 18, Battelle Memorial Institute Auditorium parking lot, Rt. 315 and King Avenue. Admission $1.00. Trunk sales $2.00 per space. Talk in on 75/15 and 52. For information: Bill, W8LLU (614) 261-7053 or Kevin, WA8OHI (614) 891-2205.

PENNSYLVANIA: The Nittany Amateur Radio Club's annual Hamfest, and now including a computer faire, July 9, Pleasant Gap Firemen's Park, Route 144, Pleasant Gap. Opens 8 AM. Tickets $3.00. Tailgaters $5.00 per space. Programs and flea market. Refreshments. Talk in on 146.16/76 and 146.25/85. For information: Dave Buckwalter, KC3CL, 1635 Circleville Rd., State College, PA 16801 (814) 234-0759.

PENNSYLVANIA: The Harrisburg Radio Amateur Club's annual Firecracker Hamfest, Monday, July 4, Shellsville VFW picnic grounds, I-81 north of Harrisburg. Shaded pavilion and tables. Admission $3.00. XYL and children free. Free tailgating. For information and table reservations: KA3HZW, 131 Livingston Street, Swatara, PA 17113 (717) 939-4957.

PENNSYLVANIA: The Murgas Amateur Radio Club (K3YTL) will hold the annual Wilkes-Barre Hamfest, Sunday, June 19, Kingston Armory, Market Street, Kingston. Doors open 6 AM for setup, 8 AM for general admission. Donation $3.00. XYLs and children under 12 free. Tailgating $1.00 extra per space. Rain or shine. Talk in on 146.61, 146.88, 224.66, 146.52, simplex. For information: Hamfest Committee, PO Box 1094, Wilkes-Barre, PA 18703 (717) 779-3882.

PENNSYLVANIA: The Milton ARC's 12th annual Hamfest, Sunday, June 12, rain or shine, Winfield Fire Co. grounds, Route 15, south of Lewisburg. Note new location. 8 AM to 5 PM. Registration $3.00. Spouse and kids free. Flea market, auction and contests. Talk in on 146.37/.97, 146.025/.625. For further details: Ken Hering, WA3IJU, RFD #1, Box 381, Allenwood, PA 17810. (717) 538-9168.

VIRGINIA: The Ole Virginia Hams ARC's ninth annual Manassas Hamfest, June 5, Prince William County Fairgrounds, Rt. 234, south of Manassas. Gates open 8 AM. Tailgating setup 7 AM. Admission $4.00. Tailgaters $3.00 extra per space. Under 12 free. Indoor exhibits, YL program, CW proficiency awards. Bring the family for an enjoyable day! For information: Bob Kelly, KA4NES, 7700 Anderson Ct., Manassas, VA 22110.

WEST VIRGINIA: The Triple States Radio Amateur Club will present its 5th annual Wheeling, WV Hamfest at Wheeling Park on Sunday, July 24, from 9 AM to 4 PM. Dealers, flea market and auction, free parking, refreshments, ARRL, SWOT booths, etc. Admission $2.00, children under 12 free. Indoor display, tables available, price of admission only but reserve space. CONTACT: TSRAC, Box 240, RD 2, Adena, OH 43901. Phone (614) 546-3930.

WISCONSIN: The South Milwaukee Amateur Radio Club's annual Swapfest, Saturday, July 9, The American Legion Post 434, 9327 South Shepard Avenue, Oak Creek. Activities being at 7 AM. Admission $3.00 per person and includes a "Happy Hour" with free beverages. Parking, picnicking, refreshments available. Also free overnite camping. Talk in on 146.94. For information and a local map: South Milwaukee ARC, Inc., PO Box 102, South Milwaukee, WI 53172.

ONTARIO: The ninth annual Ontario Hamfest, sponsored by the Burlington Amateur Radio Club, Saturday, July 9, Milton Fairgrounds. For information: Burlington ARC, PO Box 836, Burlington, Ontario L7R 3Y7.

BRITISH COLUMBIA: The Maple Ridge ARC is hosting Hamfest '83, July 2 and 3, Maple Ridge Fairgrounds, 30 miles east of Vancouver on #7 Hwy. Registration Hams $5.00; non-Hams over 12 $2.00. Food displays, swap & shop, bunny hums. Ladies' and children's programs. Plenty of camping space. Talk in on 146.20/80, 146.34/94. 20% off pre-registration. Contact: Bob Haughton, VE7BZH, Box 292, Maple Ridge, BC V2X 7G2.

OPERATING EVENTS

"Things to do..."

JUNE 11-12: The Macomb Emergency Communication Association will operate special event station KA8KTV in commemoration of Michigan tornado season. 1300Z June 11 to 2100Z June 12. 20m RTTY, 14.080 - 14.090, 2m FM 146.07/67 and the upper general class phone portion of 15, 40 and 80m. QSL with 8½ × 11 SASE to: KA8KTV, Box 291, Utica, MI 48087.

JUNE 11: The Libertyville and Mundelein Amateur Radio Society (LAMARS) will operate W9HOO near the site of the largest train robbery in the United States history from 0000Z to 2400Z. Frequencies: Phone and CW — 15 kHz up from lower 40-15m General Class band edges; Novice — 21.135 ± QRM. Certificate for large SASE to KB9BR or "Big Robbery", Box 656, Libertyville, IL 60048.

JUNE 21 - July 11: State Line ARC, Hobbs, NM, will operate station W5SZS during the World Soaring Competition, 2200 to 0400 UTC weekdays and 1200 to 0400 UTC weekends on 80 through 10 meters, 2 meters FM and SSTV. Pilots from twenty or more countries will be participating. For a special certificate send QSL info with QSO number and $1.00 to W5SZS Special Event, 209 W. Gold, Hobbs, NM 88240. SWL's are also eligible.

JUNE 18: Cape Fear ARS will operate WB4YZF from 1300-2100 Z, from the 15th annual National Hollerin' Contest. Phone 7.235. For details: Lee Brown, N4DTB, 462 Shoreline Drive, Fayetteville, NC 28301.

JUNE 10-11: The annual Bell Tower Festival in conjunction with the National Corn Throwing Contest will operate WØMLY, 1400Z to 000Z each day. Bottom 10 KC General Band. Send #10 SASE to Box 7, Rippey, IA 50235 for a nice certificate.

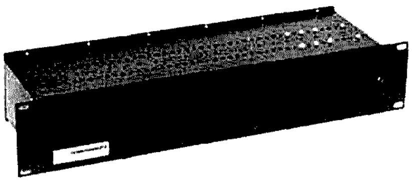
JUNE 11: The Wireless Institute of Northern Ohio (WINO) will operate special event station KO8O to commemorate Ohio Wine Week, Saturday evening 7 and 11 PM EDST on 3900 MHz and 7235 MHz. Sunday afternoon, 11 AM and 4 PM on 7235 MHz and 21360 MHz. The station will be located at an actual winery in Madison, Ohio. For a special QSL certificate send legal SASE plus 40¢ to: KO8O, 7126 Andover Drive, Mentor, Ohio 44060.

JUNE 11-13: World Communication Year RTTY Contest. Conducted by the Australian National ARTS (replacing the VK/ZL Oceania RTTY DX Contest for 1983). 0000Z-0800Z; 1600Z-2400Z; 0800Z-1600Z. Bands: 3, 5, 7, 14, 21 & 28 MHz. Classes: Single operator, multi operator and SWL. Logs of multi operators must be signed by all operators with their callsigns. Exchange: RST, Zone, Time in UTC. For more information: W.J. (Bill) Storer, VK2EG, 55 Prince Charles Rd., Frenchs Forest, 2086, N.S.W. Australia.

NEW products

community repeater panel

Ferritronics Inc. has introduced a new Community Repeater Panel, the FT124C. The FT124C is compatible with digital-coded squelch schemes, such as Digital Private Line, Digital Channel Guard, etc., in addition to the conventional EIA CTCSS tones. Up to eighteen users may be accommodated by the new panel, which includes complete audio processing, drop-out delay and time-out timer circuits on plug-in boards in a sturdy rf-shielded enclosure.



Only six connections, including power and ground, are required to interface the panel to any full duplux station, to create a very cost-effective shared repeater. For more information, contact Ferritronics, Inc., 1319 Pine Avenue, Niagara Falls, New York 14301. RS#301

antenna rotator control

The DRC-1 rotator control incorporates a crystal-controlled modem allowing all control signals to pass over standard two-wire telephone lines up to distances of five miles. With leased phone lines this distance can be increased indefinitely.

The DRC-1 control offers high versatility in azimuth control of larger antennas with shortest path-continuous rotation, or north-center 360-degree rotation, verified by a digital azimuth readout. Switch-selectable 110/220-volt operation and line-fault indicator are standard. The DRC-1 is housed for tabletop operation or can be rack mounted. The DRC-1 controls any one of three heavy-duty Hy-Gain antenna rotators with torques of 9,000, 16,000, and 23,700 inch pounds supporting vertical loads of 1,000, 7,000, and 20,000 pounds respectively.

For further information, please contact the CIM Department, Hy-Gain Division of Telex Communications, Inc., 8601 Northeast Highway 6, Lincoln, Nebraska 68505.

"To Win The World" film

"To Win The World" is a 29-minute color film that follows the crew of the million-dollar K2GL mountaintop station in Tuxedo Park, New York, during a recent DX contest. The film highlights scenes from the contest, includ-

ing the raising of a four-element 80-meter Yagi onto a 200-foot tower. It is directed and produced by Peter Bizlewicz with Larry Miller, and narrated by Bill Leonard. This film is available in 16-mm print and videotape.

"To Win The World" is a movie that will interest both ham and non-ham viewers. It's the story of a technological sport, presented in a way that captures the excitement and challenge of contesting. Films are available in three formats: 1/2-inch VHS or Beta, $75; 3/4-inch video cassette, $300; and 16-mm print, $425. The 16-mm prints are also available for rental. For more information, contact Peter Bizlewicz, 1209 Pines Lake Dr. W., Wayne, New Jersey 07470. RS#302

compact hf transceiver

The FT-77 is an extremely compact HF transceiver for active mobile or space-conscious operators. Using computer-aided design and automated insertion techniques, the FT-77 represents an advance in manufacturing efficiency and reliability. Equipped for SSB and CW operation (fm optional), the FT-77 includes digital frequency display, CW wide/narrow selection, selectable AGC, RIT, and a highly effective noise blanker.

For more information, contact Yaesu Electronics Corp., P.O. Box 49, Paramount, California 90723. RS#303

NEW products

high towers

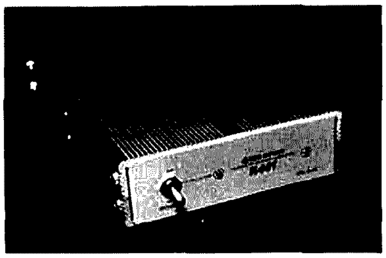
Towers up to 200 feet high can be constructed using Aluma Tower Company's all steel stack-up sections. Either 10- or 20-foot sections are fitted with connecting hardware at the factory for building-block assembly on site. All sections are welded with diagonal cross-bracing for maximum strength and minimum wind drag. The steel stacking towers are designed for permanent installation for business, civil defense, or private communications needs. Aluma Tower Company also manufactures all-steel and aluminum telescoping towers which require minimum or no extra guying.

For more information, contact Aluma Tower Company, P.O. Box 2806, 1639 Old Dixie Highway, Vero Beach, Florida 32960. RS#304

2-meter amplifier

THL's most economical new VHF linear amplifier, the new 30-watt HL-30V, is designed for use with portable 2-meter radios. The HL-30V is a high-quality, easy-to-use, 30-watt VHF amplifier. It is designed to be driven to maximum output power with only 3 watts drive from handheld radios. It will take 1-5 watts drive, with 1 watt delivering about 10 watts output. This neat little amplifier is perfect for turning

your handheld 2-meter transceiver into a powerful mobile (or base station with appropriate power supply).



The amplifier operates on 13.8 Vdc and draws approximately 4 amps maximum during transmit. It utilizes carrier operated switching (COX) with no delay and has SO-239 connectors. The HL-30V measures approximately 4 × 6 × 1 inches (100 × 158 × 30 mm) and weighs 520 grams. Suggested retail lor the HL-30V is $69.95.

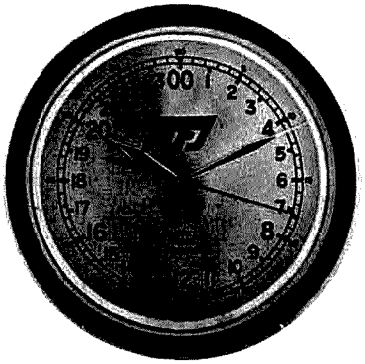
Also available fron Encomm is the HL-32V. This amplifier is similar to the HL-30V but has an FM/SSB switch to allow SSB/CW operation. It also has a high/low power switch which cuts output power by one half. Suggested retail price for the HL-32V is $89.95.

For more information, write THL Sales Department, Encomm, Inc., 2000 Ave. G, Suite 800, Plano, Texas 75074. RS#305

24-hour quartz wall clock

MFJ introduces its new 24-hour quartz-controlled wall clock. Its large 12-inch-diameter face gives excellent visibility, even across the room. This new clock is quartz controlled for accuracy to within 15 seconds a month. A sweep second hand makes precise reading easy.

A single AA battery provides over one year's operation and immunity from power line failure, and eliminates a power cord. The battery is not included.



NEW products

MFJ provides a thirty-day money back trial period. If you are not satisfied, you may return the clock for a full refund (less shipping). MFJ also provides a one-year limited warranty.

The MFJ-105 clock is available from MFJ Enterprises, Inc., for $49.95 plus shipping and handling. For more information, contact MFJ Enterprises, Inc., P.O. Box 494, Mississippi State, Mississippi 39762. RS#306

radio direction finder

BMG Engineering announces a new radio direction finder, the Super DF. Hams can use the system for sport transmitter hunts, finding stuck microphones, and hunting jammers. It's also useful for finding stuck or stolen transmitters, and search and rescue teams can use it to triangulate on boats at sea or downed aircraft.

This easy to use unit connects to any unmodified NBFM receiver (such as a scanner, hand-held, or transceiver) at the antenna input and external speaker jack. It will work on any frequency between 100 and 260 MHz with one antenna, and between 200 and 550 MHz with another antenna. This non-ambiguous system resists overloading, and neither an S-meter nor attenuator is required. All strengths of signals can be hunted. One control unit can be used with any antenna unit: base station, mobile, or hand-held. When used in mobile-in-motion, the electronics do an excellent job of averaging out reflections from nearby objects, permitting stable, accurate bearings to be taken. Not having to stop to take bearings cuts down the time required to reach the transmitter.

The SuperDF is available in kit form or assembled. Kits include plated and drilled box, drilled antenna boom, and antenna elements. The instructions include figures, diagrams, theory of operation, operating instructions, check-out and troubleshooting section, and extensive hints on hunting with the system. Construction and adjustment requires only simple hand tools, epoxy glue, and a VOM. For more information, send an SASE to: BMG Engineering, 9935 Garibaldi Avenue, Temple City, California 91780. RS#307

off-the-shelf enclosures

PacTec Corporation has released a new four-page color brochure outlining its off-the-shelf line of enclosures for the electronics and related industries. The brochure includes dimensions and prices for all injection-molded ABS enclosures, from the small, hand-held Series H enclosures to the large Series CL enclosures, measuring 12.5 × 11.6 × 8.8 inches. Enclosures for desktop and computer systems are also described.

Standard accessories such as tilt stands, wrist and shoulder straps, and cord wraps are presented, as well as a list of available design options including speaker grills, ventilation slots, and EMI/RFI shielding.

For more information, contact PacTec Corporation, Enterprise and Executive Avenues, Philadelphia, Pennsylvania 19153. RS#308

scientific instrument interference control

A new forty-page catalog (Number 831) from Electronic Specialists presents their line of instrument and computer interference control products. Protective devices for smooth instrumentation operation include equipment isolators, ac power line filter/suppressors, line voltage regulators, and ac power interrupters.

Sections describing particular scientific and computer problems and suggested solutions are included. Typical applications and uses are highlighted.

For a copy of the catalog or further information, contact Electronic Specialists, Inc., 171 South Main Street, Natick, Massachusetts 01760. RS#309

new antenna

Bilal Company has introduced a new antenna to its product line, the Isotron 15. Designed to operate on the 15-meter band, this antenna will give performance comparable to a full-size dipole in just a fraction of the space. The Isotron 15 will handle full legal power and has a bandwidth of 450 kHz, with less than 2-1 VSWR. Center frequency is adjustable for optimum performance on your favorite frequency. The antenna weighs less than 2 pounds and is just 21 inches in length. The Isotron 15 can be mounted on any 1-3/8 inch or smaller mast and in the vicinity of other Isotron antennas for a compact and unobtrusive multiband installation.

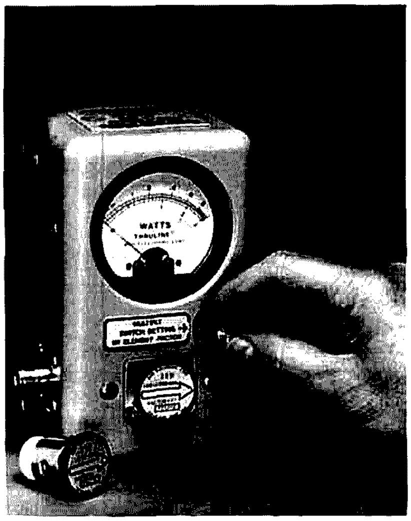
For more information, contact Bilal Antennas, Star Route 2, Eucha, Oklahoma 74342. RS#310

rf wattmeter

The new Thruline® directional wattmeter model 4410 expands the usual single full-scale power level of its plug-in element to seven overlapping power ranges.

Designed for CW and fm systems from 200 kHz to 1000 MHz and 1/4 watt to 10,000 watts, the new precision instrument uses special elements providing seven levels instead of one. The 37-dB power range covers 1/3/10/30/100/300/1000 or 10/30/100/300/1000/3000/10,000 watts with ±5 percent accuracy. Range is selected by a front-panel rotary switch which also includes a battery-level position. Elements are simply rotated for either forward or reflected power measurement.

Model 4410 Thruline® wattmeters feature



low-insertion VSWR of 1.05 or less, temperature compensation to maintain full rated accuracy from 0 to 50 degrees C, 120-percent overrange protection regardless of the selector switch position, and a choice of eighteen common rf connectors interchangeable in the field.

For additional information, contact Bird Electronic Corporation, 30303 Aurora Road, Cleveland (Solon), Ohio 44139. RS#311

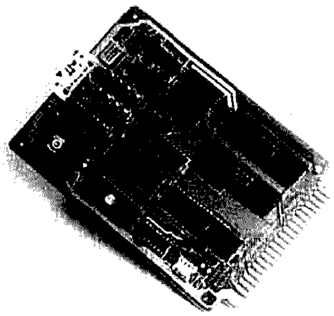
ITC-32 TT control board

Advanced Computer Controls is proud to introduce its new ITC-32 Intelligent TouchTone® Control Board. The ITC-32 addresses the need for TouchTone® control in Amateur Radio, commercial, and industrial applications with microcomputer-based flexibility and state-of-the-art Mitel tone decoding (No PLLs!).

NEW products

The ITC-32 provides twenty-eight remotely controllable logic outputs and four remotely sensed inputs. Morse code or tone encoded response messages verify command entry, and enable remote interrogation of output and input logic states. Eight of the twenty-eight outputs are buffered for high current/high voltage-drive capability, such as for direct relay drive. The other twenty outputs are TTL compatible levels. The outputs may be commanded singly, or in groups, allowing a variety of control possibilities, such as antenna direction, PL frequency, and gain controls. An additional command allows BCD programming for control of remote base frequency synthesizers.

The logic inputs may be interrogated, or may function as alarm inputs, such as for intrusion detection, over-temperature, or flood indication with external sensors. Optional connection to our telephone interface board allows landline control, and auto-dial out on alarm conditions.



For more information, contact Ed Ingber, WA6AXX, Advanced Computer Controls, 10816 Northridge Square, Cupertino, California 95014. RS#312

plastic cable wrap

The M.M. Newman Corporation introduces a multi-purpose plastic cable wrap that comes in handy packaged lengths for organizing and protecting wires, tubing, and hoses. M.M. Newman cable wrap is an expandable polyethylene cable harness that snugly grips, organizes, and protects. It is as easy to apply as tape and stays in place without adhesives or glue.

M.M. Newman cable wrap comes in three standard sizes: 1-3/8-inch diameter × 10-foot black UV resistant; a 1/4-inch diameter × 10-foot black; and an assortment pack of 1/8-inch, 1/4-inch, and 3/8-inch sizes in 5-foot lengths. Private labeling with two-color headers is also available.

For additional information, contact M.M. Newman Corporation, 7 Hawkes Street, Marblehead, Massachusetts 01945. RS#313

high-band and UHF RF amplifier

A new line of fm RF amplifiers that increase the output of low-power high band or UHF portables and mobiles from as little as 2 watts to as much as 100 watts is now available from The Antenna Specialists Co.

The compact, self-contained A/S RF amplifiers operate on 12 Vdc (negative ground) and are field-tunable across the range of 136-174 MHz in high band models or 430-512 MHz in UHF. Six high band models accept nominal RF inputs of 2, 10, or 25 watts and produce a nominal output of 25, 50, or 100 watts. The four UHF models deliver a nominal 25, 50, or 80 watt output with a nominal input of 4 watts, or 80 watts RF out with 25 watts in. Unit size varies depending on the frequency range and power output. The smallest measures only 1.56 inches high by 3.75 inches long (including connectors) and is supplied with an under-dash mounting bracket. The medium power models are 6.6 inches long by 4.6 inches wide by 2.5 inches high (including connectors) and the high power models are 2 inches longer; both have mounting flanges with four bolt holes.

For further information, contact The Antenna Specialists Co., 12435 Euclid Avenue, Cleveland, Ohio 44106. RS#314

30-meter mobile antenna

A new resonator has been added to the line of Hustler hf mobile antennas. Model RM-30 is designed for operation in the newly allocated 10.100 to 10.150 MHz band when used with a MO-1 or MO-2 foldover mast. VSWR at resonance is 1.15:1 (50 kHz under 2:1), and power handling capability is 250 watts.

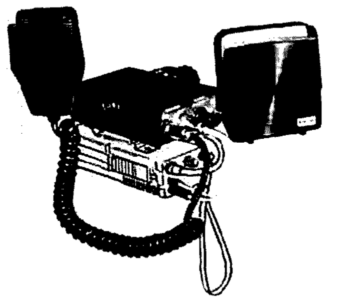
For more information, contact Hustler, Inc., 3275 North B Avenue, Kissimmee, Florida 32741. RS#315

HT conversion

VoCom announces its new HT conversion system, which makes possible the conversion of virtually any hand-held radio to full mobile operation through use of the VoCom Power Packet. The Power Packet gives 3 watts of audio output power to cover road noise, and its unique charging system keeps your HT charged and ready for portable operation. Also included is a microphone preamp to accommodate near-

ly any microphone and a hooded lamp to illuminate the HT at night. When the unit is dash-mounted, all front panel HT controls are conveniently accessible.

Smaller than many control heads, the Power Packet measures only 5 × 3-1/4 × 1-1/2 inches, thereby simplifying mounting in automobiles. The packet, external speaker, rf amplifier and the HT can all be mounted in separate locations within the vehicle, making them less conspicuous from outside the car. The packet functions as a "control head" for the system.

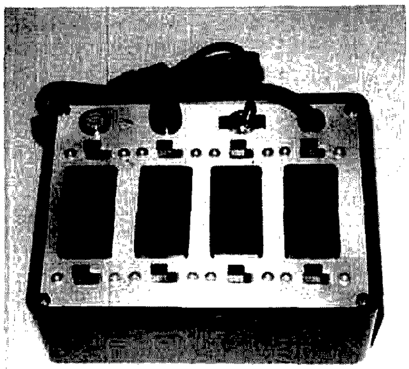


Suggested retail is $84.95; matching speaker, $19.95; external rf amplifiers are priced from $84.95. For more information, contact VoCom Products Corporation, 65 E. Palatine Road, Prospect Heights, Illinois 60070. RS#316

surge suppressor and noise filter

Kalglo Electronics Co., Inc., has added a new console-model Spike-Spiker to its existing line of voltage surge suppressors and noise filtering devices for protecting sensitive equipment from damaging voltage spikes and line noise. Called the DPC-Plus, it provides eight individually switched 120-volt, 15-ampere outlets divided into two banks of four outlets each, a main on/off switch, fuse, status lights, and 7-foot grounded heavy-duty cordset.

Voltage spikes are suppressed in six different



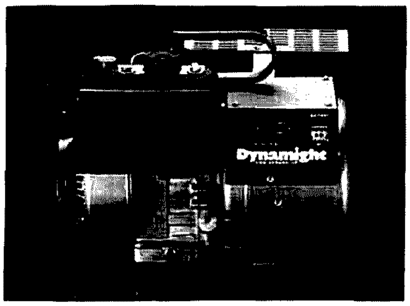
NEW products

stages on both the common and differential modes. Suppression starts at 131-volts and responds in one picosecond with an absorption capacity of 174.5 joules. Noise filtering is provided by using inductive/capacitive series — parallel lowpass networks in five stages on both common and differential modes. Separate status lights provide, at a glance, monitoring of the common and differential modes.

For more information, contact Kalglo Electronics Co., Inc., 6584 Ruch Road, E. Allen Township/Bethlehem, Pennsylvania 18017. RS#317

44-pound generator

WINCO, a division of Dyna Technology, Inc., announces the new K900 Dynamight Generator, which delivers 900 watts of ac power. The K900 weighs only 44 pounds and is UPS shippable. It features a USDA forestry-approved muffler for super-quiet running. In addition, the K900 offers automatic reset overload protection, automatic throttle control, a 76cc four-cycle Kawasaki engine and two 120-volt outlets. The K900 is manufactured in the U.S.A.



For more information, contact Walt Brammer, Sales Administration, WINCO Division of Dyna Technology, Inc., 7850 Metro Parkway, Minneapolis, Minnesota 55420. RS#318

code training programs

Three new audio cassette training programs for operators who want to increase their copying speed are available from Twin Oaks Associates.

Developed over a four-year period by Amateur Radio operators who are also mental

health professionals, the programs employ the principles of the psychology of learning to teaching code and upgrading ham skills.

The System 12 Alphabet Book© introduces basic Morse through a series of six thrity-minute cassettes. Suitable for individual or classroom use, the Alphabet Book takes students up to and beyond 7 WPM. The cost for the set is $15.

System 12©, consisting of five sixty-minute cassettes, is designed to help Novices and Technicians pass the 13 WPM FCC code test, with a speed range of 2½ to 17 WPM. The set includes a study guide and is priced at $30.

Operators already copying at 10 WPM and preparing for the Amateur Extra Class test can use System 24© to help increase their copying speed. Through a series of five one-hour cassettes, System 24© takes the operator from six to well over 30 WPM. With a study guide, System 24 sells for $30.

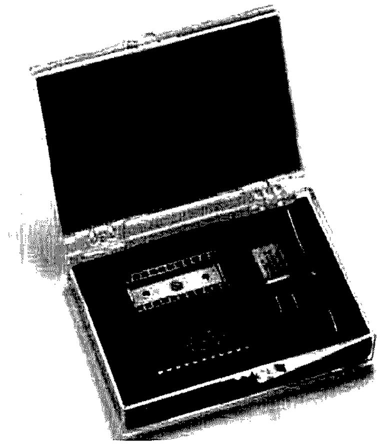
For additional information, contact Twin Oaks Associates, Route 5, Box 37, Knoxville, Iowa 50138. RS#319

DTMF receiver kit

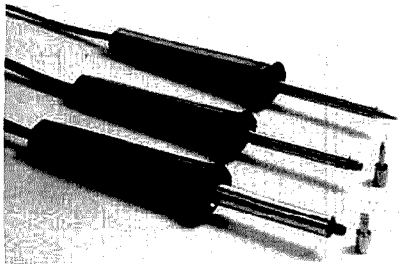
The new Teltone M-956 DTMF receiver is now available with all the parts necessary to breadboard a central-office-quality DTMF detection system. You supply only the power source.

The features and performance of the M-956 make it ideal for applications such as computer data entry, equipment remote control, telephone switching, and mobile radio. The unit's sensitivity, dynamic range, noise immunity, and low power consumption make it particularly well suited for use in communications products, and it comes in a twenty-two-pin DIP industry-standard pinout.

For more information, contact Teltone Corporation, 10801 120th Avenue Northeast, Kirkland, Washington 98033-0657. RS#320



NEW products



energy-saving soldering irons

Three new low-priced "consumer" soldering irons with advanced "Thermo-Duric" heaters were introduced by the Ungar Division of Eldon Industries, Inc., at the Electronic Distribution Show in Las Vegas.

"Thermo-Duric" heating elements reach soldering temperature faster, use less energy, last longer, and take less space than earlier wire-wound heating elements. Since the heaters were developed for industrial soldering systems, the new "consumer" line has soldering qualities and dependability appropriate for electronic technicians and serious hobbyists.

The CM-25 ($8.50), has an integral nickel-plated cone tip suitable for small and large connections. The 25-watt iron heats to 750 degrees F.

The 45-watt CM-45 and 80-watt CM-80 can use any of eleven standard Ungar screw-on tips, and have three-wire cords to prevent leakage current damage.

The CM-45 ($11.25), comes with an iron-plated pencil-tip point. Operating temperature is 700 degrees F. The large-capacity CM-80 ($17.50), comes with an iron-plated chisel tip and operates at 800 degrees F.

Slimmer, cooler handles were made possible by the more efficient "Thermo-Duric" heaters.

For further information, contact Ungar, 100 West Manville Street, Compton, California 90220. In Canada, contact Eldon Industries of Canada, Inc., 500 Esna Park Drive, Markham, Ontario L3R 1H5. RS#321

single chip repeater control

Digital Microsystems, Inc., has announced the release of a single chip repeater control (SCRC) that features crystal-controlled timing accuracy for ID, tail, and timeout timers, with the period of each timer programmable by the user. Each chip features an audio generator for generating the repeater station's call sign as well as several useful control messages such as

NEW products

"TEST" and "pF." The chips are equipped to interface directly with PL decoders and also include a PL-enable input.

Additional features include a tune input for holding the transmitter on while making adjustments, a force ID input for manual trigger of an ID sequence, user-selectable Morse code transmission rates, and compatibility with a planned autopatch controller.

With the user's call sign programmed into the chip at the factory, the SCRC sells for $89.95. An optional manual — including a data sheet and applications information — is available for $5.00.

For more information, contact Digital Microsystems, Inc., 607 Sudbury Street, Marlboro, Massachusetts 01752. RS#322

autopatch and DTMF decoder modules

Hamtronics recently released two new modules to complement their line of VHF and UHF repeaters. The autopatch module provides full telephone patch and reverse autopatch functions for a repeater or duplex rural radio telephone installation. In addition, it allows both primary repeater control via phone line *and* secondary control via the repeater receiver; it also allows a control operator to monitor the repeater receiver by telephone even when the transmitter is off. The autopatch features a choice of either automatic answer or on-air tone ringing when a party calls the reverse patch function. It also features automatic level limiting, time-out timer, tape recorder relay for logging, and access code tone muting for security. The price of the autopatch is $89.95 in kit form, $159.95 wired.

The DTMF decoder/controller module can be used with the autopatch, or can be used alone for control of repeaters and other devices by radio link. It has outputs to control two on/off functions independently. Typically, it is used to control a repeater and autopatch, but there are many other remote control jobs it can perform in radio, industrial, mining, and scientific applications. The decoder uses a four-digit DTMF code, and several safety features are provided for security against falsing or tampering. The unit is all solid-state (no relays) and uses commonly available ICs. The kit costs $89.95; the wired unit, $159.95.

For more information and complete catalog, write Hamtronics, Inc., 65F Moul Road, Hilton, New York 14468-9535. (For overseas mailing, please enclose $1 or 4 IRCs.) RS#323

technical forum

Welcome to the ham radio *Technical Forum. The purpose of this feature is to help you, the reader, find answers to your questions, and to give you a chance to answer the questions of your fellow Radio Amateurs. Do you have a question? Send it in!*

Each month our editors will select the best answer received to a question previously posed in Technical Forum. We'll send the writer a book from our Bookstore as a way of saying thanks.

noise blanker

I need a noise blanker circuit that could be used in a National HRO 500. Any suggestions? — Bill Blackwell, K8LO

pacemakers and rfi

Some time ago I saw an article about the very poor shielding of pacemaker devices used to regulate the heartbeat. It was written by a doctor in Dallas, and it should have scared any Amateur who has had a pacemaker installed. I have no idea how many, if any, hams have died because of rf getting inside a pacemaker, but recently in the cardiac section of a local hospital I was disturbed to overhear a conversation between a pacemaker recipient (an Amateur) and his doctor. The doctor did not know what my Amateur friend was talking about when he asked about the effects of rfi on the pacemaker, nor did he understand the problem when it was explained.

Here in Florida there are probably more retired hams with heart problems than anywhere else, yet I have been unable to find out how many have died as a result of pacer failures caused by rf, nor have I been able to obtain specs on any of these devices. As I understand it, the pacemaker is simply a device that emits a controlled pulse to the heart.

Do any of the readers of *ham radio* have any information on this important question? — Edwin M. Hollis, K4CN.

making verticals quieter

Bill Orr's article on vertical versus dipole antennas in the October, 1982, issue of *ham radio* verified what those of us who have used both types of antennas have found in their performance. The noisy vertical antenna can be quieted, however, without changing or modifying the vertical antenna in any way.

In order to "tame" a vertical, one technique employs a transformer at the feed point of the vertical. My first experience was with a simple quarter-wave vertical designed for 40 meters. transformer was connected as a 1:1 ratio transformer (**fig. 1**). The same The transformer consisted of eleven bifilar turns wound on the core supplied for an Amidon balun kit. The transformer can be used to match a 200-ohm load by connecting the windings as in **fig. 2**.

I used this configuration on a non-resonant antenna I built. Since this antenna has about a 200-ohm impedance, I used a 4:1 ratio transformer to bring it down to 50 ohms.

Ace Collins, K6VV, suggested this idea a few years ago when he was doing some experimental work on ground-independent antennas. The first time I tried it I thought my feed line was not connected, the antenna was so quiet! I pass this suggestion along for what it may be worth to those who favor vertical antennas because of their low radiation angle, omnidirectional characteristics, or the space limitations of small city lots. — Robert L. Crawford, WA6RYZ.

ham radio

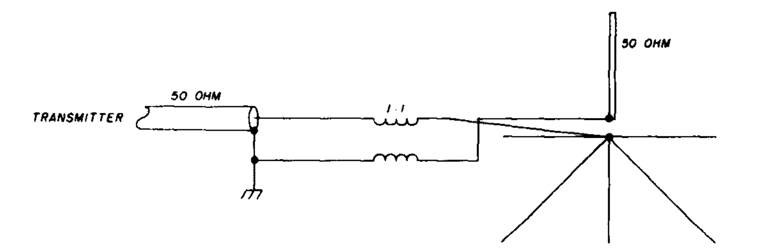


fig. 1. Bifilar 1:1 wound transformer reduces noise in 50 ohm vertical.

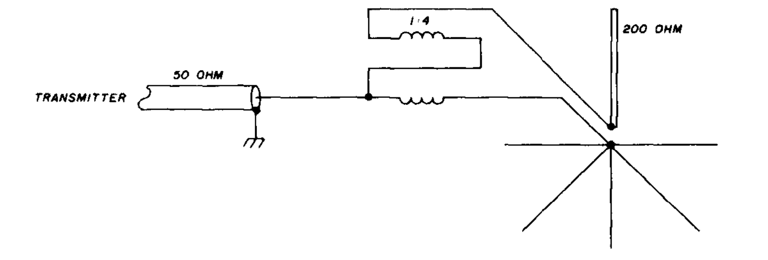


fig. 2. Bifilar 4:1 wound transformer reduces noise in 200 ohm vertical.

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ham radio magazine

Tucson's Packet Radio

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Focus
on
communications
technology

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REFLECTIONS

where were you, Idaho?

Dayton Hamvention '83 is now a thing of the past but memories of the hundreds of hams I spoke to at the show are still vivid. In looking through my logbook (and I do thank everyone who stopped by the booth, chatted with me, and signed the log*) I notice that most states were well represented (I counted 43), as were most provinces of Canada and quite a few other countries. I enjoyed seeing all of you in person, finally being able to associate faces with voices I have spoken to many times over the years.

One of the highlights occurred when I met whom I believe many of us consider the finest CW/phone operator around, a very fine gentleman, — Katashi Nose, KH6IJ. For almost three decades (that I can remember), his crisp fist has given us not only Hawaii, but also an operating style to admire and emulate. This is in addition to his many technical and personal contributions to the hobby.

Speaking of operators, it was good to see Gene (KR2N), Gary (W1EB), Dwight (W9UQO), Ted (K1OX), Rolf (KE1Y), and many other 75-meter aficionados at Dayton and commiserate on recent band conditions, discuss the latest solid-state receiver or dream about the ultimate antenna system that opens and closes the band. By the way, David (5Z4DD), thanks for making the sked at Dayton to meet one week after on 3782 — unfortunately an ionospheric storm increased attenuation on those three nights.

Besides being a great place to renew acquaintances, Dayton gave me an opportunity to discuss with you, the readers, your likes and dislikes, the type of articles you enjoy reading, specific subjects or sections that should be talked about more often or dropped. Below is just a sample of opinion expressed by the 300 or so I spoke to during the three days. Where do you fit in with your preferences? Readers told me that today's *ham radio* magazine is, in general...

... as good as the original *ham radio*

... not as good as the original

... still THE technical journal.

In regard to technical content, readers said *ham radio* is...

... not technical enough

... too technical

... just right

What do *ham radio* readers want to see? They want more construction articles, simple circuits, antenna articles; more about computer interfaces; and more coverage of VHF through microwave. They want more on RTTY, SSTV, Packet Radio, and OSCAR .... They love Bill Orr's "Ham Radio Techniques" and wonder, "Whatever happened to the DX Diary?" (Bear in mind that *every* opinion had considerable support — both pro and con.)

Some people even said they enjoy reading the editorials.

This impromptu sampling of opinion will be followed next month with a more precise evaluation of what *ham radio* readers say they want to see.

Finally, did anyone hear me say anything about the fleamarket? Here goes. It's a four letter word that starts with R and ends with N. It rained again. Too bad — Dayton Hamvention's reputation was, to a large extent, built on its flea market. Wait till next year.

Rich Rosen, K2RR
Editor-in-Chief

*Here's a random sample of some of the people I spoke to at the show:

KC4EG	WB1ADY	W8PJ	KR2N	W8TP	VE5ZV	K6IR
WD4C	WB9NDF	W8IMF	W1EB	G3ZAY	W7CFX	HL9KT
KA0APK	W4LQX	XE2UAA	W9UQO	WA0QNZ	W6HPH	W4ETO
WA9JWL	WB3JZO	N8DRI	K1OX	KH6KD	JA2AQO	N8CGG
KA9OTX	W0TN	5Z4DD	VE3APG	VE3MFO	JA2CEJ	W6SUI
K4KJ	WB8MVR	K8SS	KH6IJ	WB8GJP	N9CZK	W7FFP
W5XQ	K8UCN	N3BM	WA2LQQ	W3LPL	K1FI	CP1PO
K2GHV	VE3NGI	KB0PM	K9HI	W8LRL	OA4HJ	KA4QVK
WD8IKV	N9BW	VE5XK	N2NY	DK4EI	KA7NKX	W0ZO
WB2ABD	K1CKS	KE1Y	WB6QMD	W0RAN	N2CF	W5DFU
WA2AOG	AF9B	K2BQ	N4DKE	KB2M	XE1LL	WB9IVR

presstop de W9JUV

PCB CONTAMINATED DUMMY LOADS MAY POSE A SERIOUS HEALTH HAZARD in many ham shacks! According to the Center for Disease Control in Atlanta, many RF dummy loads manufactured as recently as the late 70s utilized transformer cooling oil containing PCBs, which have been linked with liver cancer. PCB use is now prohibited by law, and all contact with any oil that could contain PCB should be avoided. Even fumes from a warm load could be dangerous in a poorly ventilated shack! Area EPA offices may have disposal suggestions.

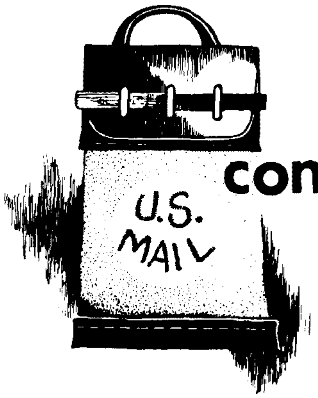
EXPANSION OF THE 10-METER REPEATER SUBBAND DOWN TO 29.0 MHZ was proposed by the FCC at its May 12 meeting. Ten meter repeaters are presently restricted to 29.5 to 29.7 MHz, with 100 kHz offset the accepted standard. Under this new proposal, PR Docket 83-485, Amateur satellite downlinks at 29.3-29.5 MHz would become subject to FM repeater interference.
10-Meter Simplex Interference To Satellite Users Has Become a significant problem in the past few years. Increased use of 10-meter FM has driven FM users to below 29.5 to find clear frequencies, while more and more SSB operators have moved above 29 MHz for the same reason. The resulting interference to 29.3-29.5 MHz satellite downlink signals has become a major problem at times, and even triggered some on-the-air confrontations.
Without Suggesting Any Solutions To This Problem the FCC asked that it be one of the factors considered by Comment submitters. The comment due date was pending at press time.
The 28.3 MHz Lower Phone Limit Proposed In The FCC's further NPRM on the phone band expansion, PR Docket 82-83, is also being questioned by some 10 meter users. A world-wide system of beacons now operate between 28.2 and 28.3 MHz, but setting the lower U.S. limit at the top of the beacon band would push some foreign SSB operators into the midst of the beacons. Perhaps a 28.4 MHz lower limit would be better, leaving 28.3-28.4 MHz open for foreign phone operation. Comments close July 1, with Reply Comments open until August 1.

HAND-HELD RADIOS VS. USER HEALTH HAS BECOME a legal issue in New Jersey. A fire chief there has sued General Electric, alleging his use of one of their hand-helds over a 14-year period damaged his sight and hearing. At issue is GE's alleged negligence in not providing a warning of possible health hazards, despite a recommendation by the federal government in 1973 that such a warning be provided with portable transceivers.
Whether Close Exposure To Moderate RF Fields actually causes physical ailments has been the subject of heated debate for years. Despite many government and industry studies no clear-cut consensus has been reached. Attempts have been made on the local level, most recently in Massachusetts, to closely regulate all transmitter operators, and an on-going effort (strongly supported by ARRL's Biological Effects of RF Energy Committee) is being made for the adoption of a federal preemption law with exemptions for Amateur Radio.
The Effects Of A Decision Favoring The Fire Chief could have an even more serious effect on Amateur Radio than the current antenna ordinance problems, barring federal preemption. Local governments, acting to protect citizens, could enact legislation that would severely restrict if not bar operation of Amateur transmitters within their borders.

VOLUNTEER ADMINISTERED AMATEUR EXAMS WERE "AN ABSOLUTE, UNQUALIFIED SUCCESS" at this year's Dayton Hamvention, according to the FCC's John Johnston. With only one FCC staff member present to act as "overseer," Dayton Amateur Radio Association members were able to administer 683 exams to 484 applicants. The volunteers were obviously well prepared for their task, as the program came off extremely smoothly despite only one evening of "formal training" with the FCC. Oddly enough, the ARRL quietly made a last minute attempt to scuttle the Dayton exam session, on the grounds that it was likely to be improperly done and would thus set the entire volunteer program back!
Proposed Questions For The Volunteer Exam Program are already being sought informally from the Amateur community, even though the exam program itself is still to be acted on by the Commission. It's felt that having a pool of appropriate questions on hand would facilitate preparation when the FCC is ready to move on both the overall exam program, PR Docket 83-27, and the Novice "No-Mail-Back" proposal, PR Docket 82-727. Action on the latter could take place as early as June.

BURBANK (ILLINOIS) TOWER CASE MOVED CLOSER TO COURT after a magistrate recommended to the presiding judge that the city's motion to dismiss the Amateurs' suit be denied. In his recommendation the magistrate agreed that the Amateurs' argument that their constitutional rights of free speech and civil rights were both violated by Burbank's anti-tower ordinance raised a federal issue, so the case did belong in U.S. District Court.
A Status Call Has Been Set By The Judge for June 21, when he's expected to adopt the magistrate's recommendation. A date for the trial should be set soon after that.
ARRL Funding Of Amateur Radio Legal Cases will essentially cease, following a vote to that effect at the April 21-22 League Directors' meeting. The League will, however, continue to offer other forms of support to Amateurs with legal problems and may, under special circumstances, offer financial assistance as well.

PHASE III-B COULD BE IN SPACE BY THE TIME THIS SEES PRINT with a June 16 launch date announced at press time. An AMSAT crew was to leave momentarily for the French Guiana launch site for final checkout and fueling. If Phase III-B is up, check with ARRL or an AMSAT net for status, as it's not to be used until completion of the post-launch test.
OSCAR 8 Is Now On Mode J Only, due to ongoing battery problems with the aging bird.



comments

briefcase Bobtail

Dear HR:

Just a few comments in regards to W6BCX's article on the Bobtail curtain featured in the February and March issues of *ham radio*.

I received word just four days before departure for Haiti that I had a license waiting for me. What to do about an antenna for 20 meters? I had just reviewed the few paragraphs on the Bobtail curtain in the *ARRL Antenna Book* when W6BCX's timely article came along. This was enough for me to make preparations to build such an array.

During the last couple of days before I was to leave, I built up a parallel tuned network consisting of a 70 pF variable capacitor and 12 turns of No. 14 wire spaced 1 inch apart. Ten turns tapped one turn up from cold end and about 3/4 maximum capacity gave a perfect match into 2000 ohms. So with a roll of No. 18 copperweld wire, some insulators made from 1/2 × 2 inch pieces of 1/4 inch phenolic, my Swan dual meter SWR bridge, the parallel tuned network built into a 4-1/2 × 5 × 2-1/2 inch aluminum box and a few short lengths of RG-58 stuffed into my briefcase, I was off. Destination, 120 miles west of Port au Prince in the mountains of the panhandle of western Haiti.

It took me, with help from my son, about 20 minutes to build the antenna. Twenty feet of bamboo put the northeast leg 24 inches above the new galvanized metal roof of his carport. The center leg also was 24 inches above the metal of the back porch roof and just six feet above my proposed operating position. The southwest leg was about three feet from the ground on a sloping hillside of about 45 degrees.

Results were outstanding. From deep in a mountain valley, with a ridge all across the north from east to west, 300 to 500 feet higher and a quarter to half a mile away, I worked all areas of the United States. I received 59+ reports from my home country of northwest Wyoming and southern Montana and 59+20 from the Denver area. The rig was HH2DR's TS520, sometimes operating on battery power. I worked a CN8 off the northeast end of the antenna just before the ARRL DX contest and had an OE6 and an HA6 call me during the contest even though I was not contesting. They gave me 55 to 57 reports. I probably had some distortion of the signal because of the large mango trees near both end elements. Compared to the 2-element quad of HH6BG located just 100 yards north, whom I had worked quite a few times from my home QTH, it was 2 to 4 S-units better. It was not the fault of the quad but instead of its location. The mountain hillside is 200-300 feet higher, begins 50 feet directly in front of the quad and at a 45 degree angle.

I brought the antenna home and will be using it on Field Day. By fall I hope to have a 40 meter version up and aimed at Europe and the South Pacific. I'm looking forward to some good 40 meter DX next winter. A match for my 20 meter 5-element log-Yagi at 70 feet it is not, but truly a fine, easy-to-pack antenna with gain broadside and rejection off the ends.

Thanks to W6BCX for his research.

Paul M. Rich, WA7BPO
Cody, Wyoming

power supply

Dear HR:

In the March, 1983, article "Dual Voltage Power Supply," the LM317 could be replaced with a 723-type regulator IC realizing the following benefits: lower cost, current limiting features, and, what I view as the most important improvement over any LM317 series pass transistor design, improved voltage regulation. An additional benefit could be improved ripple rejection.

The only drawback is an increase in circuit complexity required to accommodate the feedback and the internal voltage reference. The 723 has enough output current to drive the existing pass transistor. The 723 is available at Radio Shack with required specs and circuits for about 89 cents.

Peter J. Schuch, WB2UAQ
Little Ferry, New Jersey

noise figure data

Dear HR:

I was rather surprised at some of the noise figure data presented by Dennis Mitchell, K8UR, in the article "GaAs FET Performance Evaluation and Preamplifier Application" in *ham radio*'s March issue, and I would like to present some additional information regarding the performance of the Mitsubishi devices tested by Mr. Mitchell.

At the 1982 meeting of the Central States VHF Society, at Baton Rouge, Louisiana, there was a preamplifier noise figure competition. These tests

were conducted with the current Hewlett-Packard programmable automatic noise figure meter with matching noise head. The results, however, departed significantly from the figurés quoted in the article, particularly for the MGF-1200s.

Here are some of the results:

device	noise figure (dB)	frequency
MGF-1200	0.27	144
MGF-1200	0.42	144
MGF-1402	0.42	144
MGF-1200	0.48	144
MGF-1200	0.38	220
MGF-1402	0.39	220
MGF-1402	0.45	220
MGF-1200	0.47	220
MGF-1402	0.49	220
MGF-1402	0.40	432
MGF-1402	0.58	432

The Central States VHF Society results were significantly better than those of the author for the MGF-1200 at 144 MHz. Assuming that Mr. Mitchell presented median noise figure values in his article, then the figures presented above are at least 0.1 dB better in the worst case, taking the stated ±0.23 dB root-sum of squares uncertainty into account. in the best case for the MGF-1200 at 144 MHz the deviation from the author's noise figure is 0.3 dB!

The figures for the MGF-1402 are included to show that this device seems to reach a plateau at 432 MHz, and is not really a cost effective device at 144 MHz, with most GaAs FET users preferring the MGF-1200 or other similarly priced device at lower frequencies. Finally, the price structure that is mentioned in the article is about one year out of date, with the MGF-1200 currently selling for around $10, rather than the $15 indicated, and the MGF-1402 available for $15 or less, as opposed to $30.

From my experience, anyone using the MGF-1200 at 144 MHz should expect, and get, substantially better results than those indicated by Mr. Mitchell, in terms of noise figure attainable.

Jack C. Parker, KC0W
Bismarck, North Dakota

Amateur packet radio: part 1

The history and operation of packet radio are examined along with its requirements for software and hardware

Imagine sitting down *in front of your station for an evening. You get out your 2-meter fm transceiver, attach it to a cable coming from an 8 × 8 × 3-inch "black box" connected to your data terminal. After turning everything on and initiating a short dialog between the terminal and the box, you enter a friend's call letters. After a short pause you see:*

****CONNECTED to (call sign)*

on your terminal. From this point on, everything you type appears on your friend's terminal, and everything he types appears on yours. Your friend could be within simplex range, or within voice repeater distance, or accessible only via a series of linking stations. In fact, you might need a satellite link to talk to your friend!

He asks, "Would you like a copy of my latest program for playing 'Escape The Maze'?"

"Sure," you reply, "only my compiler can't handle your gigantic programs. Why don't you just send me a dump of the machine language (binary) program?"

"No problem. Let me know when you're ready," he sends back.

You go over to your home computer, power it up, load your communications program, connect it to the box instead of the terminal, and type, "OK, let 'er rip."

Then you start your file-loading program and wait. Soon, binary data begins arriving from your friend at slightly less than 120 bytes of data per second. You sit back relaxed, knowing that even though the QSO is being held under noisy conditions, with occasional QRM breaking through, you won't receive a single bit incorrectly.

After the program has been stored away, you resume your conversation. It is almost boringly error-free, and with the speaker disconnected from your radio you don't even hear the QSO, which is being periodically interrupted by the automatic identification of both stations in CW. Later on you try out the new program and, sure enough, find you've received the whole thing perfectly.

Does this sound like magic? It shouldn't — it's happening right now with packet radio.

Packet radio promises to open new worlds of communications undreamed of just a few years ago by making possible the rapid transfer of digital information over great distances — with a virtual guarantee of integrity down to the last bit. This is tremen-

By Margaret Morrison, KV7D, and Dan Morrison, KV7B, 4301 E. Holmes, Tucson, Arizona 85711

dously attractive. Not only can traffic be exchanged between hams equipped with data terminals, but just as easily between a ham and a computer, or between two computers.

Let's look first at what a packet is and then at the history of packet communications and the kind of hardware and software packet radio requires. We will use the two most familiar systems to serve as examples, although others are in use as well. These two are the VADCG (Vancouver Amateur Digital Communications Group) system and the TAPR (Tucson Amateur Packet Radio) system.

what is a packet?

Packet radio is a relatively new form of digital communications. It has some characteristics in common with older forms, such as ASCII and RTTY, now both familiar to the Amateur community. In all of these modes information is coded in binary form, that is, as a series of 1s and 0s. The information is translated into an audio signal consisting of alternations between two tones, and the audio signal then used to modulate an rf signal to produce an FSK or AFSK transmission.

In an ASCII or RTTY system, the transmission typically consists of a sequence of individual characters separated by periods of unmodulated carrier transmission. In order for the receiving station to interpret the characters correctly, extra transitions are added at the beginning and end of each character (start and stop bits). Depending on reception conditions, anywhere from all the information to virtually none of it may be received correctly; what's not received correctly may be garbled or missed completely.

A packet consists of binary data (which might be ASCII, Baudot, or some other code), and the modulation techniques may be essentially the same as for conventional ASCII or RTTY, although the exact interpretation of the tones may be different. The VADCG and TAPR TNCs produce AFSK, but more sophisticated schemes are being developed. (The TNC, or terminal node controller, is the "black box" referred to in the introduction to this article. It is a complete microcomputer-based communications system with a good-sized memory, 30 kilobytes in the case of the TAPR TNC. It does all the work involved in sending and receiving packets).

In a packet, the individual characters, or bytes, are run together with no space at all between. This eliminates the need for both the start and stop bits as well as the dead time between characters. The result is much more efficient information transfer. The analog of start and stop bits are sent only for the beginning and end of the packet, and the transmitter is keyed only while information is actually being sent.

Extra information is inserted into each packet that enables the receiving station to determine automatically whether the packet was received without error. Thus every correctly received transmission is acknowledged. The sending station can keep retransmitting its information until it is assured that it has gotten through. Other features of the packet which facilitate this "handshaking" are described later.

history of packet radio

Packet switching is a technology that was developed to tie computer users into a network which could extend over a wide area. It has been used for many years over common carrier lines, both commercially and by government. The first large-scale packet network in North America was ARPANET, set up in 1969 by Bolt Beranek and Newman, Inc., for the Defense Advanced Research Projects Agency. This network introduced packet switching, in which each message sent is broken up into small packets and each is switched to its destination over the quickest communications path available at that instant. Data interconnections are typically 50-kilobit-per-second wideband lines, and the packets are passed from node to node until they arrive at their destination. Typical end-to-end times are 250 milliseconds, and receipt of data is acknowledged.

Other networks around the world soon began operation, and today there are many government and commercial computer networks, such as TYMNET and TELENET, which allow users all over the country to access thousands of computers remotely.[1]

Packet radio experiments began in the 1970s. One of the largest packet radio systems, based at the University of Hawaii and known as the ALOHANET, linked together a number of computers and users, and also provided access into ARPANET and satellite links.[2] Other systems were developed for the purpose of providing distributed automatic digital communications for remote sensing stations.

Packet switching networks (both wire and radio based) generally use one of two methods for routing packets from the originating station, through intermediaries, to the destination. In one system used by TYMNET and others, a central controller determines the optimum path for a particular pair of stations on the basis of the stations present in the network at any time. In the other system, the network itself is intelligent and determines the routing between stations. This is the system that was pioneered by ARPANET.

North American Amateurs first entered the picture in Canada, where, beginning in 1978, the Department of Communications encouraged the use of packet radio by permitting Amateur packet transmis-

sions and by giving exclusive use of 221 to 223 MHz and 433 to 434 MHz to packet and digital transmissions. Taking advantage of this ruling, VADCG, a group in Vancouver, British Columbia, designed the first well-known Amateur packet radio TNC, and soon TNCs became widely distributed.[3] Their use in the U.S. followed a rule by the FCC making such ASCII transmissions legal in March of 1980. Finally, in October of 1982, the FCC revised Part 97.69, lifting many restrictions on digital communications and advanced data transmission. Today many experimenters using the VADCG TNC, the TAPR TNC, and homebrew systems are hard at work, developing this new mode of communications.

anatomy of a packet

The basic element in packet radio is the frame — a string of bits with a specific format. The bits are presented to the transmitter on a modulator output line. In the case of the TAPR and VADCG TNCs, the modulation system uses 1200-Hz and 2200-Hz tones and coherent (phase-continuous) FSK, with a data rate of up to 1200 bits per second; it is compatible with the Bell 202 standard modem. Other modulation systems being developed for Amateur use include minimum shift keying (MSK), and various forms of phase shift keying (PSK). These schemes, which are more efficient than ordinary FSK, are useful for long-haul traffic, especially via satellite.[4]

The FSK signal is related to the bit stream according to specific digital encoding rules. The most commonly used system is non-return to zero inverted (NRZI) encoding. In this system, a transition from one tone to the other is interpreted as a 0, whereas no transition during the bit period is a 1. Such a method is used because, according to the rules by which the frame is constructed, a transition is guaranteed at least once in every five bit periods. This is needed to keep the receiving station in "sync" with the incoming data.

The actual structure of the frame varies from one packet radio system to another. The structure makes possible, among other things, the delivery of the message to the proper recipient and a system for ensuring data integrity. The most frequently encountered format for frames is known as HDLC, or High Level Data Link Control. Each HDLC frame consists of six fields, as shown in **fig. 1**.

FLAG1	ADDR	CONTROL	DATA	FCS	FLAG2

fig. 1. Schematic representation of an HDLC frame.

In order of transmission, FLAG1 is first. It is at least eight bits long, consisting of the bit pattern 01111110. This particular combination is unique to FLAG1 and FLAG2, and appears nowhere else in the frame. Part of the transmitting station's job is to alter the message content of the frame to prevent this combination from appearing elsewhere (a process known as bit-stuffing). This alteration is, of course, undone by the receiving station. FLAG1 (which may be repeated several times before the rest of the frame is sent) says, "Get ready! Here comes a frame!"

The ADDR (address) field varies among the various packet radio systems developed in the Amateur community. HDLC requires only that it be at least one byte long. It typically contains the source address, and may contain the destination address and perhaps routing information. The address field contains the information which permits delivery of the packet.

The CONTROL field also varies among systems. The length of this field specified by HDLC is one or two bytes. The information contained in this field typically includes acknowledgment information for previous packets successfully received; an indication that the sender would like to begin talking (connect) to the destination station; a request to terminate the conversation (disconnect); or other "supervisory" functions, such as requests to stop transmitting or to resume transmitting (referred to as flow control).

The DATA field consists of zero or more bytes of information (zero in the case of simple acknowledgments, for example). They may be in any bit pattern — ASCII characters, part of a binary program, you name it. (The FCC, however, would like you to have available enough information so they can decipher your data!) The HDLC standard requires that when five consecutive 1s appear a 0 be inserted. This is the bit-stuffing mentioned above. It prevents data from being mistaken for flags, and also ensures frequent tone transitions if NRZI encoding is used. Upon reception, these extra 0s are discarded. Typically, the maximum data length is 128 to 256 bytes.

The last item in the frame prior to the ending flag bits is the FCS, or frame check sequence, an extremely important two-byte number computed by the transmitting station based on all the bits in the frame following FLAG1. If the frame is received in garbled condition it is extremely unlikely that it would be garbled in such a way as to produce the same FCS. The FCS is separately computed by the receiving station and, if both numbers agree, there is virtual certainty that the frame was received as sent.

Finally, the frame ends with another byte of flag field, thus indicating to the receiving station that the previous two bytes were indeed the FCS.

protocols

What we have described is not yet truly packet radio. It could be called "frame radio," the exchange of frames of information. The protocol, in addition to specifying the structure of the frame, determines the contents of the ADDRESS, CONTROL, and possibly the DATA fields. It also determines action to be taken in various situations. For example, just exactly what should be done if the first, second, and fourth frames received in a single transmission check out, but the third does not? Or, what should be done if the other station suddenly stops responding? The list of "what-ifs" increases rapidly as other users join the frequency.

The interchange of packets results in communications between the participating stations on more than one level. The ISO, International Standards Organization, has defined a model network structure consisting of seven "layers." The first three, levels 1, 2, and 3, are concerned with communications and are the ones of interest to us. Each consists of a set of related tasks which would ordinarily be handled by correspondingly related processes (electrical or software). The ISO layer structure does not define the specific protocol to be followed to accomplish the tasks of any level, and the operation of each level should be independent of how lower-level tasks are performed.[5]

Furthermore, each layer is "transparent" to the levels above it. This means, for example, that information used to direct actions by a level 3 process is treated as data by the level 2 process. A packet is structured like an onion. Each process peels off the applicable control information before passing the remainder to the next higher level.

The bottom layer is called the physical layer. It is concerned with such things as modulation and transmission techniques, signaling the beginning and end of packets, bit-stuffing, and maintaining synchronization with the incoming data stream. The second level, or data link layer, defines the use made of the address, control, and FCS fields of the packet. Level 2 is responsible for setting up and maintaining a connection or data link with the other station. This includes verifying data integrity, acknowledging receipt of intact frames, retransmitting unacknowledged frames, and performing various link control functions. The third level, the network layer, defines routing functions and inter-network communication. Level 3 is concerned with setting up and maintaining routing tables for communication between stations which are not in direct contact. Amateur packet radio has implemented some level 3 functions but not all.

An additional set of rules, a collision avoidance protocol, is necessary for packet radio but not for communications over wires. Since stations cannot receive at the same time they are transmitting, "collisions" occur when two or more stations transmit simultaneously. A scheme for avoiding repeated collisions must ensure different retransmission times after an initial transmission has failed. If all stations can hear each other, as is the case when all transmissions are made on the same frequency and all stations are close together, all that is needed is to impose a short random wait time for stations retransmitting a packet. If a central controller (or a satellite) transmits on one frequency and listens for all other transmissions on another frequency, a more elaborate scheme is required.

The HDLC frame structure described above is imposed on levels 1 and 2 of all protocols implemented so far for Amateur packet radio, and both the VADCG and TAPR TNCs use LSI chips that perform many of the level 1 and 2 tasks. The two most widely used protocols, VADCG and AX.25, are thus functionally equivalent on level 1 and quite similar on level 2.[6,7] AX.25 is modeled on X.25, a standard developed by the Consultative Committee for International Telegraph and Telephone (CCITT) of the ITU[8]. AX.25 was put forward by a group of Amateurs at the AMSAT packet conference in October of 1982. AX.25 specifies the address as containing Amateur call signs of both the sending and receiving stations, with optional routing information in the form of the call signs of stations requested to relay, or digipeat, the packet. The VADCG address field contains a numeric address of the sending station only; packets setting up the connection contain call sign information in the data field. Relay by an unspecified digipeater can be requested. The control functions implemented in AX.25 are summarized in **table 1**. Most control func-

table 1. Level 2 control functions.

RR	Receive ready: acknowledge receipt of information frames by specifying the sequence number of the last packet received.
RNR	Receive not ready: request to stop sending (receive buffers full).
REJ	Request retransmission of missed frames after receipt of a frame number larger than expected.
DM	Disconnected mode: response to a packet other than a connect request.
SABM	Set asynchronous balanced mode. This is a connect request.
DISC	Disconnect request.
UA	Unsequenced acknowledgment: sent in response to a connect or disconnect request.
FRMR	Reports an abnormal condition; that is, receipt of a packet with an undefined or invalid control byte.

tions can be performed by a packet which also transmits data. Fewer level 2 control functions are specified in the VADCG protocol.

implementation

If you have a home computer, you are probably wondering where you can get a packet radio program for it. You may even be thinking about writing one yourself. The only hitch here is that you need more than a program. At a minimum, you need some hardware to enable the computer to control the radio push-to-talk line, put signals into the microphone input, and interpret signals on the speaker output. Specialized hardware, such as an HDLC controller, is very desirable. This hardware must be able to generate interrupt requests to the computer. The program itself should take care of the input and output requirements of both the radio and the terminal through interrupt processing. You can't afford to miss part of an incoming packet because you got busy parsing a line from the terminal! This means that the program probably has to be written at least partly in assembly language. Interpreted languages, such as BASIC, are commonly used on small computers, but they are neither fast enough nor versatile enough for real-time programming of this kind. These obstacles are not insurmountable, and in fact many hams have been successfully running packet radio programs on various home computers.

There are disadvantages with this approach, however. These programs are not very portable: they work on a specific computer with a specific operating system, and depend on the specific configuration of the hardware "extras." The programming has to be done over for each different type of computer. Modifying a protocol would be a major undertaking involving reprogramming many computers. Furthermore, many hams who don't own computers or who don't want to get involved in a programming project are interested in packet radio. After all, an RTTY terminal unit or a CW keyboard need not be connected to a computer. This is why most Amateurs involved in packet radio are using a terminal node controller. The TAPR and VADCG TNCs have standard terminal interface connections, and provisions for versatile radio interfaces. The ROM memory chips can be programmed with software implementing a standard packet radio protocol, and, once such software is written, it can be easily transferred to any similar TNC. Since the TNC is basically a dedicated microprocessor, the demands of radio communications do not interfere with a resident operating system.

packet radio — communications of the future

Hams all over North America are now involved in sending packet radio messages across town on VHF on UHF bands. Digipeater relays and ordinary voice repeaters make it possible to communicate over distances of 100 miles or more. Packet radio mailboxes and bulletin boards are on the air in several areas. Interest is growing rapidly in this newest mode of communications. With more experimentally inclined packeteers joining the ranks, exciting developments will be forthcoming. The emphasis for the future will be on long-distance communications and inter-network linking protocols. Experimental hf packet communications has been done on 10 meters. Inter-network communications through UHF and microwave linking stations using high data rate modulation techniques is envisioned. The digital special communications channel on the AMSAT Phase III-B satellite will see use by packet radio stations. Groups are working on protocol standards for this application and on L-band amplifiers to allow inexpensive access to this satellite mode. Possibly the most ambitious project in the works is a packet radio satellite with a store-and-forward mailbox as well as direct relay capability.

Part two will continue with a detailed description of the TAPR terminal node controller; it will provide a clearly defined set of interface requirements and point out pitfalls to be avoided in making reliable radio connections.

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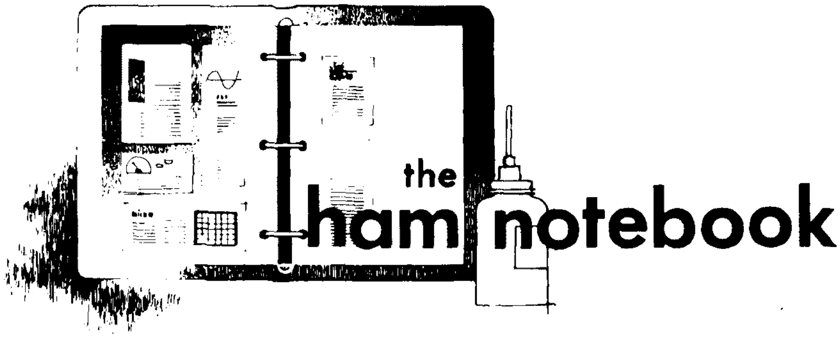
Second ARRL Amateur Radio Computer Networking Conference Proceedings, March 19, 1983. This recent publication contains descriptions of packet radio systems, including implementation details.

Tucson Amateur Packet Radio Corporation Packet System Beta Test (1983). This manual contains information on AX.25, VADCG protocol, modulation, and HDLC.

ham radio



heatsink cooling fan

Most modern transceivers in the 200 watt class mount the amplifier on the back of the cabinet. The heatsink is exposed, and should be cooled by a breeze. A muffin fan is just right to make that breeze.

Surplus houses sell them for around $12, but you can pick them up new at ham flea markets for around $4. These fans run from 120 Vac, but are fast and noisy, masking the speaker signal. To slow the fan down, put a 600 ohm 20/30 watt resistor in series with the 120 volt line. I put mine in front of the breeze to keep it cool. A four-inch resistor will mount on clips attached to the holes in the fan used for mounting (see **fig. 1**).

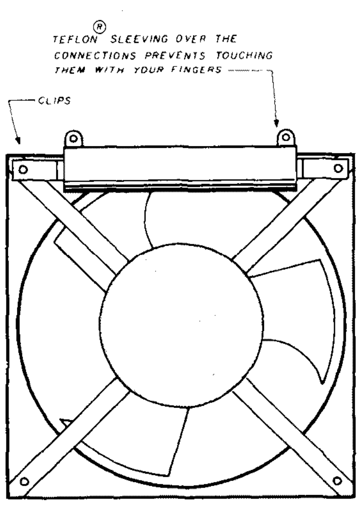


fig. 1. Heat sink cooling fan.

Ed Marriner, W6XM

HF antenna

Some time ago I tried a 160-meter antenna described in *Editors & Engineers Radio Handbook* by Bill Orr, W6SAI, (21st Ed., Section 27-17, fig. 22). The results were quite gratifying, probably because of the high ground conductivity under the antenna. The ground for the antenna was at the base of a 40-foot TV tower.

I now have a small home at the seashore on a small lot, too small to put up a 120-foot dipole for 75 meters. In the past I had tried a single-wire 1/4-wave antenna, but with only limited success. Then this year I put up the one shown in **fig. 2**. I first put up the 75-meter portion, made with 300-ohm TV ribbon to the specs given in the *Handbook*. My results on 75 were much better than with the 1/4-wave dipole, but this antenna, of course, worked on only one band.

Next I tried using two lengths of 300-ohm ribbon, cut for 40 meters and 20 meters, and slung under the 75-meter section. Because of the close coupling to the 75-meter section, these did not work. But it was interesting to note that the performance of the 75-meter antenna was not affected by the addition of these two sections. I replaced the 40-meter and 20-meter sections with wire, to form a 1/4-wave antenna on these bands. Now all three antennas tuned up well. VSWR at 3.825 MHz was 1.4, at 7180 it was 1.2, at 14275 it was 1.4, and at 21.300 it was 1.4. Normally it would not be necessary to use an antenna tuner, but with the TS-120S solid-state transceiver, maximum output occurs at only 50 ohms. Also, by using the tuner I work over the full portion of these phone bands.

construction

The spacers were made from three plastic clothes hangers purchased at the local discount store for 97 cents. Each hanger was cut up to get the straight sections. Six were cut to 9-inch lengths and these were used for the 40-meter and 20-meter sections. Four were cut to 6-inch lengths for the outer supports of the 40-meter section. Holes were drilled for passing the wires through them, and then the wire was tied to the supports with a piece of fishline. See **fig. 3** for details.

Here I might remind you to make sure the grounded portion of the SO-239 cable connector is secured to the tower base with a strap or heavy wire (#14 or larger). The one grounded side of the 300-ohm ribbon is soldered to the SO-239 casing and the other three wires are soldered to the center pin. After soldering, the SO-239 was coated with Dow-Corning DC-9 for weather protection. Connection to the equipment is by means of thirty-five feet of RG8/U.

For use on a small lot, this system seems to work quite well, and it has a high angle of radiation, which I prefer for contacts up to 800 miles on 75. Don't expect this type of antenna to compete with a high half-wave antenna on any of these bands, but it does perform well for reasonable distances — even with its short length.

J.F. Sterner, W2GQK

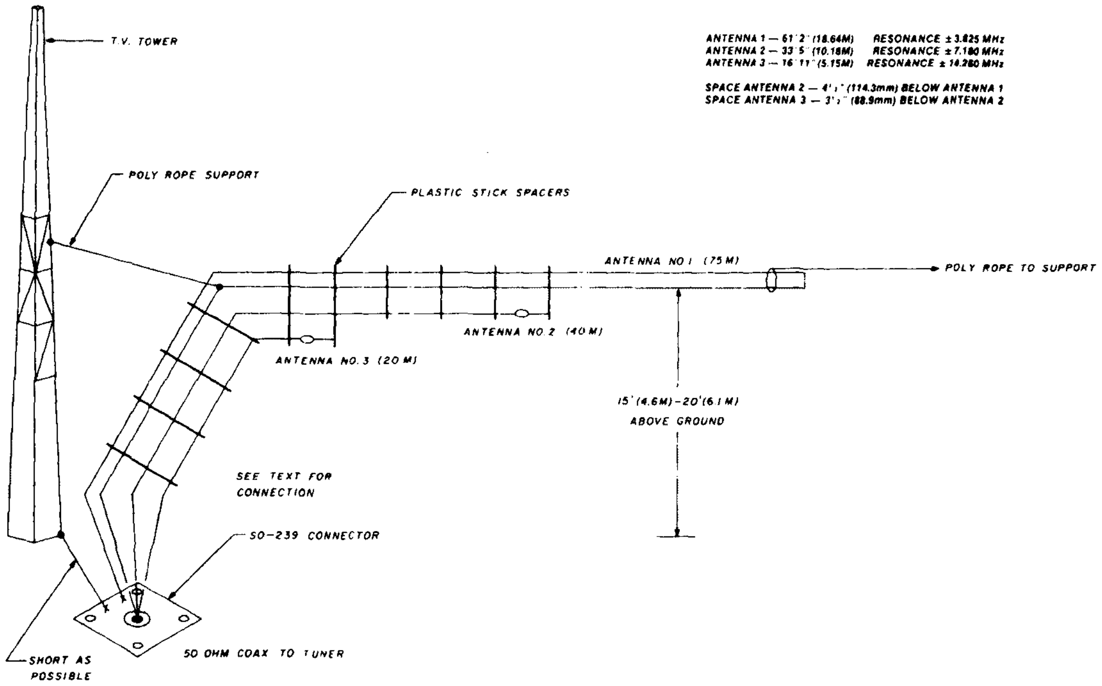


fig. 2. Construction details of the antenna for 75 through 15 meters.

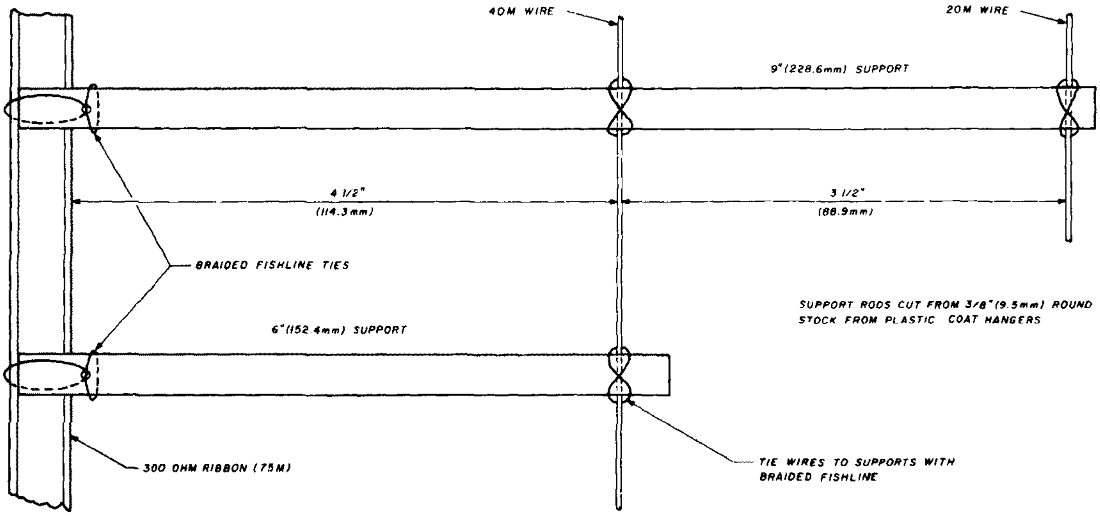


fig. 3. Support sketch.

vertical phased arrays:
part 3

Array impedances, measurements, and calculations

This is the third in a series of articles on phased vertical arrays by K2BT.

In Part 1 (May), the author examined essential design parameters, and more importantly, the *assumptions* underlying design. (In the past, incorrect assumptions have misled designers into constructing less than optimally performing arrays.)

Part 2 (June) continued with relative power plots of two- to four-element arrays indicating the correlation between physical and electrical (phase) spacing and performance.

This month, K2BT's discussion includes the determination of self- and mutual-impedances, the importance of an extensive ground system, and a tabulation of mutual and driving point impedance values for some popular vertical phased arrays. — **Editor**.

In Part 2[1] various types of arrays were examined and relative power (in dB) plots were shown. We saw how specific physical arrangements of elements, current amplitude ratios, and phase displacements formed beams. By varying current amplitude ratios and phases, the forward beam width or the rejection characteristics of a given physical array were modified. The question now is how can these drive conditions be created in a real array? To do this we need information about element impedances in order to design the feed network.

Knowledge of self-impedance and mutual impedances, as well as factors that influence them, is essential because everything will be either directly or indirectly affected by these parameters.

self-impedance

The self-impedance of an antenna at any frequency is a function of the element length, its radius, ground plane loss, and coupling with other nearby antennas. Strictly speaking, the last two items are not components of self-impedance. However, when measuring self-impedance, both may be present in the reading of *apparent* self-impedance.

Although resonant elements are not required for an array, their use simplifies calculations and provides the following advantages:

1. An *open-circuited* 1/4-wavelength element presents virtually no coupling. This simplifies measurement procedure and ensures best conditions for accuracy of self- and mutual impedance readings.

2. The resistive component of self-impedance is normally higher than ground loss resistance which results in reasonable efficiency.

3. Ground plane evaluations and comparisons are easier to make because more information is available about the 1/4-wavelength resonant antenna than about other types of vertical antennas.

element length and radius

An article on Yagi design by James Lawson, W2PV,[2] provides data on the relationship between an element's resonant length and its radius. (When using this source, be sure to refer to error corrections[3]) It's important to use a full wavelength when calculating length-to-radius ratio, K, for W2PV's equations. For determining parameters of a resonant grounded 1/4-wavelength element, I have revised W2PV's chart as shown in **fig. 1**. In the Yagi antenna

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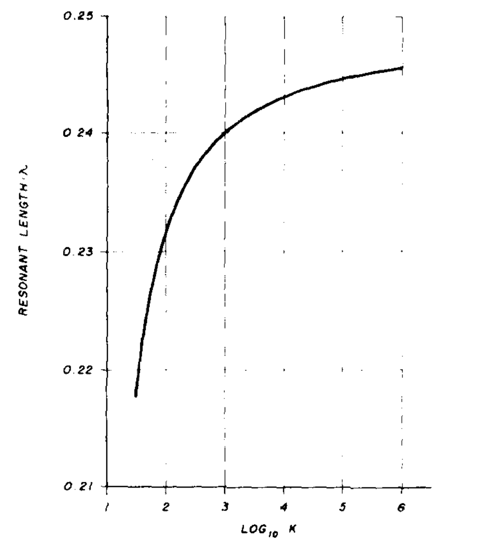

fig. 1. Resonant length of a quarter-wavelength grounded element as a function of k (operating wavelength to element radius ratio).

design, emphasis was placed on the reactance component of self-impedance, ignoring the effect that radius has upon the resistive component. In an all-elements-driven array as compared to a parasitic array, it is more important to know this effect. A review of the Amateur literature yields a range of values for a 1/4-wavelength vertical resistive component of impedance; these values are probably all correct. Any disparity is probably due to the different antenna element diameters that are used. The theoretical self-impedance of a physical 1/4-wavelength high vertical is 36.5 + j21[4] which assumes the use of an infinitely conducting ground plane and an infinitely thin element. Obviously neither of these conditions is physically realizable. However, even if an infinitely thin element could be used, it still would have to be shortened to achieve resonance — and in so doing the resistive component would decrease. A real element, having real thickness, would reduce resistance some more since it requires a further reduction in length in order to achieve resonance. Kraus[5] shows that l/r ratios in the range of 60 to 1000 are equal to a resistance variation from 34 to 36 ohms, with 35 ohms as an average value. He uses an element's actual length when calculating l/r. The comparable data for *reactance* change compiled by W2PV would show a variation for K from 240 to 4000. When resistance is plotted against the logarithm of K, we see a virtually straight line, showing a slow reduction in resistance as the element diameter is varied from 1.5 to 24 inches.

ground planes

Considerable controversy surrounds the subject of required ground plane size and its influence on antenna performance. The ground plane essentially establishes an image antenna to represent the other half of a dipole. The better that image, the lower the ground loss and the lower the radiation angle. How large the ground plane should be is answered by examining the near field (within the first 1/2 wavelength), and far field (to at least 6 wavelengths) components. The near field requirements for proper pattern formation is satisfied by a ground system composed of wire radials; a sufficient quantity allows us to get quite close to the theoretical resistance. At the lower frequencies the far field usually must be left to nature, since it would be prohibitively expensive to provide so large a radial wire or mesh ground system. Even the large a-m broadcast antennas are located in salt marshes whenever available to take advantage of the high conductivity of earth for many wavelengths beyond the reach of the radials.

My experience correlates closely with the work reported by Jerry Sevick, W2FMI.[6,7] His graph of resistance versus number of radials used on 40 meters is applicable for 80 meters as well. I used radials averaging 0.3 wavelength in length, composd of PVC No. 24 hookup wire, and laid them on the ground. The only difference noted is that resistance did not decrease as rapidly as his graph shows. For instance, I never found resistance below 40 ohms with 40 radials, but at 60 radials and greater the data correlated more closely. This discrepancy is probably attributable to the differences in soil conductivity; the land under my array is part of a moraine, and consequently represents very low conductivity earth. *All* indications are that with *120* 1/4-wavelength radials, resistance of a resonant 1/4-wavelength vertical is within a half ohm of the theoretical value regardless of the underlying soil conductivity. Another effect I noticed which W2FMI did not comment upon was that as radials were added, the element length had to be slightly but continually adjusted upward to maintain resonance.

coupling with other antennas

The attempt to approach the theoretical self-impedance value can be frustrated by inadvertent coupling of the antenna under test to another antenna. As will be seen when discussing mutual impedance, the effects are subtle and can be easily mistaken for ground plane differences. These effects can go in both directions — you may think you are achieving theoretical self-impedance with a 30-radial ground plane, or conversely that a 120-radial ground plane has several ohms loss. If you encounter either of

these indications, suspect coupling with another antenna (or something acting like one even if you don't "see" it). Another indication of this problem is a significant departure (at 80 meters — several inches) in element length for resonance. I had a tower guy wire (adequately broken up with insulators, I thought) whose lowest section ran to an anchor at the base of a tree. This section was approximately 1/4 wavelength and it found sufficient ground conductivity in the tree roots to present lossy coupling to one of my array elements. Though I knew that element wasn't right, I could not see anything that would act as a resonant antenna around it. That guy wire didn't look as if it had a ground plane! The solution was to insulate it at the anchor, thus decoupling the section of guy wire.

I am sure many Amateurs will identify with this frustrating experience: the first element of a multielement array is erected and adjusted for resonance. The length is carefully recorded and the second erected. Then, letting the first element remain connected to its feed cable, the second element is checked for resonance, found too long, and is readjusted downward. Reconnecting the second element to its feeder, the first element is now found too long. And so it continues; the result is that the elements end up considerably shortened below their uncoupled resonant length. This is mutual coupling at work and the error was in failing to open-circuit other elements when making self-impedance measurements. Other elements, at or near resonance and within about 0.35 wavelength of the antenna being measured, will manifest inductive coupling. Unless you're aware of what is happening, you may diagnose this inductive reactance to be due to the element's being too long. Shortening it will bring it to "resonance" and this may be accompanied by a satisfactory reduction in resistance (perhaps even below theoretical), but all this changes when the second element is open-circuited. It is well to remember that this situation can also occur inadvertently with a conductor not recognized as acting as an antenna. However, as we shall soon see, this same effect — mutual coupling — is the very same process used to advantage to create field enhancement and cancellation in arrays.

mutual impedance

Coupling between elements is a function of element lengths, distance between elements, relative attitudes of elements (e.g., parallel, co-linear, echelon), and ground plane losses. Ground losses are not actually a component of theoretical mutual impedance but in a practical situation they become a part of the *apparent* mutual impedance. (Mutual impedance is a term that relates to the interaction of two

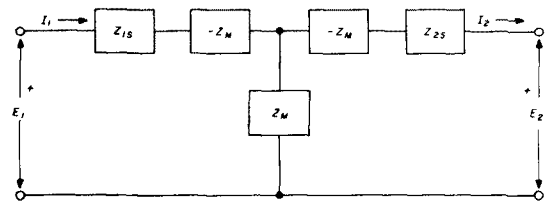

fig.2. Four-terminal equivalent network for two verticals.

or more antennas which are close enough to each other to cause their driving impedances to be different from their self-impedances.) The unit of measurement — ohms — may be, like any impedance, resistive or reactive, or both. Such antennas are coupled by an impedance which appears to be in common with all elements. (Driving point impedance calculations only require the mutual impedance between pairs — that is, two elements at a time be measured.) Mutual impedance between antennas is similar to mutual inductance between coupled coils; the impedance relationship can be both depicted and its value measured in the same way. In **fig. 2** the driving point impedance Z_1 or Z_2 of each vertical as measured at either set of terminals reacts to the presence of the other vertical as though its self-impedance Z_{11} or Z_{22} had a common impedance Z_{12} in series with it. Z_{12} is, by definition:

$$Z_{12} = -E_2/I_1$$

Although useful mathematically, it doesn't provide a practical basis for measurement. The voltage and current relationships existing in a system of antenna elements, each mutually coupled to one another, have the same form as the voltage and current in a general network. Writing their mesh equations produces:

$$E_1 = I_1Z_{11} + I_1Z_{12} + \ldots + I_nZ_{1n}$$

$$E_2 = I_1Z_{21} + I_2Z_{22} + \ldots + I_nZ_{2n}$$

$$\vdots$$

$$E_n = I_1Z_{n1} + I_2Z_{n2} + \ldots + I_nZ_{nn}$$

where $E_1, E_2 \ldots, E_n$ are voltages applied to elements 1,2,...N

$I_1, I_2 \ldots, I_n$ are element drive currents

$Z_{11}, Z_{22} \ldots, Z_{nn}$ are element self-impedances

Z_{12}, Z_{21} ..., Z_{1n}, Z_{2n} are mutual impedances and are denoted by dual subscripts which are always different. As in general networks, mutual impedances with the same subscripts but with reversed positions, (e.g., Z_{12} and Z_{21}), describe the identical impedance (from the Reciprocity Theorem).

If the equation for each drive voltage is divided by that element's drive current, the following driving point impedance terms are obtained:

$$Z_1 = E_1/I_1 = Z_{11} + I_2Z_{12}/I_1 + \ldots + I_nZ_{1n}/I_1 \quad (1)$$

$$Z_n = E_n/I_n = I_1Z_{n1}/I_n + I_2Z_{n2}/I_n + \ldots + Z_{nn}$$

Notice that each element's driving point impedance consists of its self-impedance and includes terms for the mutual impedances between it and each of the other elements. The *influence* of the mutual impedances upon the driving point impedance is a function of the drive currents (amplitude and phase) to other elements. Although at first glance these equations appear quite formidable and look like there are too many unknowns for solution, this is not the case. Having selected an array configuration and the driving current ratios and displacements for the field plot, we already know what the currents need to be.[1] If we could find a way to reduce the complexity and consequently the number of unknowns, a means for deriving mutual impedances might be devised. Fortunately there is one. Since each mutual impedance we need to know exists between only two elements, we can write a simpler set of equations:

$$E_1 = I_1Z_{11} + I_2Z_{12}$$

$$E_2 = I_1Z_{12} + I_2Z_{22}$$

If the terminal of element 2 is connected to its ground plane, the drive voltage E_2 becomes zero and:

$$\begin{aligned} E_1 &= I_1Z_{11} + I_2Z_{12} \\ O &= I_1Z_{12} + I_2Z_{22} \end{aligned} \quad (2)$$

Solving for the driving point impedance yields:

$$Z_1 = E_1/I_1 = Z_{11} - (Z_{12})^2/Z_{22}$$

and solving for the mutual impedance Z_{12} gives

$$Z_{12} = \pm \sqrt{Z_{22}}\,(Z_{11} - Z_1) \quad (3)$$

Note that all references to voltages and currents have been eliminated. We are now in a position to find all the remaining unknowns.

mutual impedance measurement

Provided the elements are 1/4 wavelength or less, the procedure is: open-circuit all elements; measure the self-impedance of element 1; connect element 2 terminal to its ground plane; measure the driving point impedance of element 1; and open-circuit element 2.

If there are additional elements, connect element 3 terminal to its ground plane; measure the driving point impedance of element 1; and open-circuit element 3.

Following the same sequence, all remaining elements are measured from element 1. When completed, a similar set of measurements are taken from element 2, starting with self-impedance and then measuring the various pairs of driving point impedances, and so on with each remaining element. This procedure allows each element to be individually treated as the reference element of each pair of elements for mutual impedance measurements. When completed, the same mutual impedance will have been read from each side of every pair. This provides a check on previously determined calculations. I am continually amazed (even though I know it is supposed to happen) by the close coincidence of the resulting value for mutual impedance as determined from either element of a pair! This occurs, as it should, even when the two self-impedances are quite *different*.

using 1/2-wavelength elements

What if the elements are significantly longer than 1/4 wavelength, specifically a 1/2 wavelength? Open-circuiting these elements from the ground plane will not decouple them (in all likelihood, coupling will be found to increase if the length is exactly a 1/2-wavelength). Means for temporarily sectioning other elements into two electrically separate halves must be provided so that self-impedances are measured with the temporary sectioning reconnected and *that element* connected to its ground plane. I have no experience with this situation but I believe the array can be driven properly, provided the high impedance at the bases of the elements can be handled.

In antenna texts, mutuals are always referred to current loops (maximum current points). Mutuals derived from measurements as above are referred to the base of the elements. These are quite different values, just as self-impedances differ according to whether they are measured at a voltage or current loop.

mutual impedance calculations

Data is taken from a 40-meter 4-square array with elements spaced 0.272 wavelength at 7.0 MHz. The elements are not alike, not resonant, and the ground plane is quite lossy. Data are shown for two elements and mutual coupling was measured from each.

table 1. List of mutual resistance and reactance between two physical 1/4-wavelength verticals separated by 0 through 1.5 wavelength spacings.

spacing	R	X	spacing	R	X
0	+36.57	+21.27	.80	−9.25	+6.13
.05	+35.83	+12.14	.85	−6.66	+8.15
.10	+33.67	+3.77	.90	−3.75	+9.28
.15	+30.22	−3.55	.95	−.78	+9.50
.20	+25.70	−9.59	1.00	+2.00	+8.87
.25	+20.40	−14.18	1.05	+4.38	+7.52
.30	+14.63	−17.22	1.10	+6.16	+5.61
.35	+8.75	−18.71	1.15	+7.26	+3.36
.40	+3.11	−18.72	1.20	+7.63	+0.97
.45	−1.99	−17.39	1.25	+7.28	−1.33
.50	−6.27	−14.97	1.30	+6.30	−3.35
.55	−9.53	−11.71	1.35	+4.81	−4.92
.60	−11.66	−7.94	1.40	+2.99	−5.94
.65	−12.61	−3.97	1.45	+1.00	−6.35
.70	−12.43	−0.13	1.50	−.94	−6.15
.75	−11.25	+3.32			

Equation 3 is used to calculate the mutual impedance.

Measurements from Element A (referenced as Element #1)

Element A	Z_{11}	$= 45.73 + j\,8.19$	Self-impedance of A
Element B	Z_{22}	$= 42.53 + j\,5.72$	Self-impedance of B
Element A	Z_1	$= 46.98 + j15.66$	Driving point impedance of A with B grounded
	Z_{12}	$= 12.53 - j12.95$	Calculated mutual impedance

Measurements from Element B (referenced as Element #1)

Element B	Z_{11}	$= 42.53 + j\,5.72$	Self-impedance of B
Element A	Z_{22}	$= 45.73 + j\,8.19$	Self-impedance of A
Element B	Z_1	$= 44.20 + j12.79$	Driving point impedance of B with A grounded
	Z_{12}	$= 12.63 - j13.34$	Calculated mutual impedance

Note the following:

1. There is a nomenclature interchange for the self-impedances of the elements, denoting the change in reference element for the measurement of mutual coupling.

2. There is only a small increase in resistive component when measuring the effect of coupling, requiring a highly accurate impedance bridge.[8]

3. At this spacing, the effect of coupling is decidedly inductive on the measured element.

4. There is reasonably good correspondence in the mutual impedance calculation from either side of the pair of elements, despite the differences in the individual elements.

5. The measured mutual impedance is quite different from theoretical values. (See **table 2.**)

As a further verification of measurements and calculations, this test is useful and instructive: With element 2 connected to its ground plane, drive element 1 from a 50 to 100 watt source while measuring current at the terminals of each element. The ratio of the current flowing in element 2 to element 1 is equal to the ratio of the mutual impedance to element 2 self-impedance:

$$I_2/I_1 = -Z_{12}/Z_{22}$$

(This identity is a rearrangement of eq. **2.**)

Since ratios are involved, the only restraint on the current measuring device is that it be linear. Although phase angles *are difficult to measure* when the reference points are located at some distance, current amplitudes can be measured and this identity is useful as a verification of impedance measurements and calculations, even if only the magnitude of the mutual impedance vector can be obtained. When performing this test, if there are more elements, open circuit them. If driving with more than 50 watts be careful of those open-circuited elements; don't provide a ground return through your body. You may be surprised to find how much energy is being coupled.

The calculations for mutual impedances require a square root extraction. Which sign to use? As general guidance, the polar vector angle of the root is *always* lagging except at spacings less than about 0.15 wavelengths. For a specific calculation the pattern of sign changes seen in published sources is an aid. Mutual resistance and reactance vary with element separation in the nature of a damped sine wave, starting with both signs positive at zero separation and proceeding through cyclic sign variations

thereafter. For example, suppose at 1/4-wavelength separation with 1/4-wavelength elements your calculator or computer produces the square root extraction $-13.7 + j15.1$ (polar notation $20.4 \quad +132.2°$). The polar angle shows lead and it should be lagging. Looking at published sources we see confirmation for this. Subtracting 180° from the polar vector angle will produce the correct signs for resistance and reactance. To aid in determining signs I have converted the table of mutual resistances and reactances shown by W2PV, to grounded physical 1/4 wavelength values in **table 1**.

The question arises: "Why bother measuring mutual impedances? Why not use published values from antenna texts?" The best answer is another question: "Why not also use textbook values for self-impedance?" Most Amateurs measure self-impedance because they want to be sure the element length is resonant at the frequency of interest or because they know from experience that the actual self-impedance can differ considerably from the theoretical value. Theoretical mutual impedance derivations are quite complex and solutions often use different simplifying assumptions. The result is that few textbook sources — except those which obtained data from a common origin — agree exactly. Regardless of source, the following assumptions apply: infinitely conducting ground plane; infinitely thin element; and element lengths measured in physical wavelengths. Element radius has a relatively small effect on mutuals. The element length assumption can be determined from the values for zero separation between elements (see first line in **table 1**). This is the self-impedance of a single element and may be recognized as identical with theoretical self-impedance. (Applies to equal length element data only.) For example, the value $36.5 + j21$ means that physical 1/4-wavelength elements had been assumed. The length difference (over resonant length) will not seriously affect driving point impedance calculations, but the assumption of lossless self-impedances will. **Table 2** lists mutual impedance between 1/4 wavelength high elements from several sources compared to an average of 16 measurements I have made.

table 2. Values of mutual impedance between two quarter-wavelength high verticals. Data from five different sources. (Gehrke's entry represents *measured* data for a real vertical over a real ground.)

source	mutual impedance (0.272 spacing)	mutual impedance (0.385 spacing)
Brown	17.49 −j17.01	2.96 −j18.47
Jasik	17.47 −j16.01	6.00 −j17.50
Jordan	17.55 −j16.37	1.66 −j18.99
Mushiake	17.51 −j15.70	4.80 −j18.75
Gehrke	13.20 −j16.24	0.20 −j16.61

The resistive component differs most. Despite these differences, if no means of measurement is available, there is something to be said for using theoretical values; at least there is recognition they exist rather than ignoring them entirely. However, as I have previously emphasized, the significance of deviation from optimum drive conditions increases with the complexity of the array. When I first became aware of the need to take mutual impedances into account for the feed network, I used theoretical values. There was improvement in F/B, but it was still far from what is achievable.

You may have wondered if an element driving-point impedance could have a negative resistive component, and if so, what that means. This is entirely possible with arrays of more than two elements, particularly with close spaced arrays or arrays employing non-unity current ratios. Elements exhibiting this condition are being driven by energy coupled from other elements; instead of *receiving* any drive from its feeder, this element is *sending* drive back into the feed network. This is still a coupled passive system, *in equilibrium, merely* observing the law of conservation of energy.

calculations of driving-point impedances

Using **equation 1**, I have calculated and listed in **table 3** the driving-point impedances of several arrays discussed in **Part 2** using measured mutuals. (For smaller spacings, values were estimated based on extrapolations of my data). For a comparison, the 4-square array driven impedances are also calculated using mutual impedances from **table 1**.

Data common to all calculations:

Element effective radius = 0.7 inch
Element height = 62.7 feet
Self-impedance = $36.4 + j0$ ohms
Frequency = 3.8 MHz

notes and comments

1. The 3 element in-line and the 1/8-wavelength 4-square have elements which exhibit substantial negative resistance components in their driving point impedances.

2. Nearly all driving point impedances show substantial reactance, requiring some care in establishing correct phasing.

3. All arrays except one exhibit unlike driving impedances, *ruling out equal power* distribution networks where *equal current* amplitude is intended.

4. Note the difference in driving point impedances in

table 3. Mutual and driving point impedance values for some popular vertical phased arrays.

array	current ratio	mutual impedances	driving point impedances
2-element, λ/4 spacing*	1/1; 0°, −90°	Z_{12} = 15 −j15	Z_1 = 21.4 − j15 Z_2 = 51.4 + j15
3-element in-line, λ/4 spacing	1/2/1; 0°, −90°, −180°	Z_{12} = Z_{23} = 15 −j15 Z_{13} = −9 −j13	Z_1 = −6.6 − j21 Z_2 = 51.4 + j0 Z_3 = 79.4 − j39
2-element, λ/2 spacing	1/1; 0°, −180°	Z_{12} = −9 −j13	Z_1 = 45.4 + j13 Z_2 = 45.4 + j13
triangular array, 0.289λ spacing	1/0.5/0.5; 0°, −90°, −90°	Z_{12} = Z_{23} = Z_{13} = 10 −j16	Z_1 = 28.4 − j10 Z_2 = 78.4 + j4 Z_3 = 78.4 + j4
4-square array, λ/4 spacing	1/1/1/1; 0°, −90°, −90°, −180°	Z_{12} = Z_{13} = Z_{24} = Z_{34} = 15 − j15; Z_{14} = Z_{23} = 3 − j17.5	Z_1 = 3.4 − j12.5 Z_2 = 39.4 − j17.5 Z_3 = 39.4 − j17.5 Z_4 = 63.4 + j47.5
4-square array, λ/4 spacing (using table 1 mutual impedance data)	1/1/1/1; 0°, −90°, −90°, −180°	Z_{12} = Z_{13} = Z_{24} = Z_{34} = 20.4 − j14.18; Z_{14} = Z_{23} = 8.41 − j18.72	Z_1 = −0.37 − j22.08 Z_2 = 44.81 − j18.72 Z_3 = 44.81 − j18.72 Z_4 = 56.35 + j59.52
2 × 2 array of arrays, λ/4 spacing	1/1/1/1; 0°, 0°, −90°, −90°	Z_{12} = Z_{13} = Z_{24} = Z_{34} = 15 − j15, Z_{14} = Z_{23} = 3 −j17.5	Z_1 = 18.9 − j33 Z_2 = 18.9 − j33 Z_3 = 83.9 + j3 Z_4 = 83.9 + j3
4-square array, λ/8 spacing	1/1/1/1; 0°, −135°, −135°, −270°	Z_{12} = Z_{13} = Z_{24} = Z_{34} = = 30 − j3, Z_{14} = Z_{23} = 25 − j9	Z_1 = −1.27 − j13.18 Z_2 = 18.97 − j4.76 Z_3 = 18.97 − j4.76 Z_4 = − 10.78 + j21.67

*This 2-element, 1/4-wavelength spaced array is probably the most common phased array configuration used by Amateurs today. Please note that the driving point impedances are different.
Editor.

the 1/4 wavelength spaced 4-square using actual mutual impedances as compared to the use of theoretical values. Current and phases in the latter case will not occur as intended in a real array.

5. Note the 2 element 1/2 wavelength spaced array (not shown in Part 2). Because of the equal driving impedances, here is one of the few instances of an array which operates as intended regardless of feeder length, as long as they are equal and a 1/2 wavelength delay line is inserted in series with one of them. Except for VSWR, Z_0 of coax is not important. The antenna pattern in this case is *not* a function of the coaxial cables Z_0 (characteristic impedance) though the VSWR still is.

We tend to become accustomed to thinking of an antenna, just as any discrete component, as having a fixed impedance at any frequency. The concept that elements within an array present impedances that are determined by other element drive currents (amplitude and phase) is, at first, difficult to appreciate. That these impedances may have negative components of resistance also can be a bit unsettling. Yet when an array is looked at mathematically as a general network which includes the impedance branches represented by mutual impedances, these seemingly unusual effects can be seen to be physical realities. Consequently, the rest of this coupled system, the feed network, must be designed for these driving impedances as the terminations.

If we expect to switch directions with this array, then we need to be sure that each physical element presents the same driving point impedance appropriate to the electrical position in the array it is assuming. I have found that equalizing self-impedances is the best means for doing this. Each element is adjusted for length to present the identical reactance (if resonance is the objective, then this is zero reactance). Assuming all elements have the same radius, radials are added to those elements showing higher resistive components. At the 100 radial level, it is not unusual for a spread of +20 radials to occur among the ground planes of the elements in this effort at equalization.

summary

We have worked our way through the design of vertical phased array antennas. A number of typical arrays were examined, as well as the current requirements of each element and the driving point impedances that must exist to cause the array to operate as designed. What remains is to design the feed network which will create conditions as they must appear, not at the element terminals, but at the end of the feed lines coming from those terminals. By now you are aware, if you weren't already, that feed lines are an integral part of the feed network.

There is no unique network which achieves the necessary current amplitude ratios and phase displacements. We can get to that objective in a number of different ways. In the next article the design task will be of use A,B,C,D parameters in single matrices as a tool. If this technique is new to you, I believe you will find this approach most interesting. You will see that this is a powerful and versatile means of network design, useful not just for antenna arrays, but for other network applications as well.

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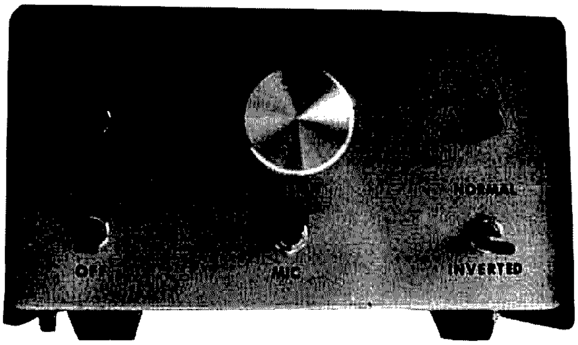
4. Prof. John D. Kraus, W8JK, *Antennas*, McGraw-Hill Book Publishing Co., page 262.
5. Prof. John D. Kraus, W8JK, *Antennas*, McGraw-Hill Publishing Co., page 245.
6. Jerry Sevick, W2FMI, "The W2FMI Ground-Mounted Short Vertical," *QST*, March 1973, page 13.
7. Jerry Sevick, W2FMI, "Short Ground-Radial Systems for Short Verticals," *QST*, April, 1978, page 30.
8. Forrest Gehrke, K2BT, "A Precision Noise Bridge," *ham radio*, March, 1983, page 50.

In commenting on vertical phased arrays, several writers have cautioned against placing arrays near trees. The apparent assumption is that trees represent resonant loss elements or somehow disturb the field so that the radiated pattern will be changed. I remain unconvinced. At wavelengths 40 meters and longer, I have measured self- and mutual impedances of elements, among trees, at all seasons of the year without seeing any significant changes that are not also seen on a pair of 40-meter elements completely away from trees. Small variations (0.3 to 0.5 ohms) are seen in self-impedances, depending upon soil moisture conditions, which are reflected in mutual impedance measurements. Since all elements are affected in the same way, these small changes cannot significantly affect radiation patterns. Examination of published mutual impedance data indicates that the presence of conductive elements, *resonant or not*, within about 0.1 wavelength of an element will significantly affect mutual impedance in unanticipated ways. Prudence would therefore dictate that nothing conductive, or even partially so, which could act as an antenna be allowed within that distance. If despite this precaution array patterns are indeed disturbed, my advice is to look for something that may be acting as a real conductive antenna in the immediate area of the array, or to re-evaluate the feed network. — K2BT

ham radio



RTTY and the Atari™ computer

Turn your Atari home computer into an RTTY terminal for either Baudot or ASCII

If there's one area in Amateur Radio that is becoming dominated by microprocessors, it's certainly RTTY. It's now common to find an RTTY operator using either a home computer or a piece of commercial gear fully dedicated to RTTY. RTTY is basically a digital form of communications, and as such it lends itself well to the use of computers. Applying a computer to RTTY requires that some basic problems first be solved. This article describes those problems and shows how they are solved in the process of making an Atari computer into an RTTY terminal (**fig. 1**).

basic problems: receiving and transmitting

When you tune your receiver to a ham RTTY station, you hear an alternation of two tones, called a low tone pair, which consist of a 1275-Hz "mark" and a 1445-Hz "space." The duration of these tones determines the character speed, measured in words per minute. A device called a terminal unit receives the two tones and generates a voltage-on state when mark is present and a voltage-off state when space is present (see **fig. 2**).

It's here that the serial interface to the computer comes into play. The serial interface detects the start of the pulse string, all on/off voltage transitions, and the end of the pulse string coming from the terminal unit. A pulse string represents a single character, and is stored as a binary number in a holding register in the interface. The computer reads this binary number and processes it before the next character appears in the serial interface. Processing usually means printing the character on a CRT, TV screen, or LED display.

The terminal unit designed for this application is shown in **fig. 3**. It is a receive-only device whose operation is controlled by the XR2211 chip. The resis-

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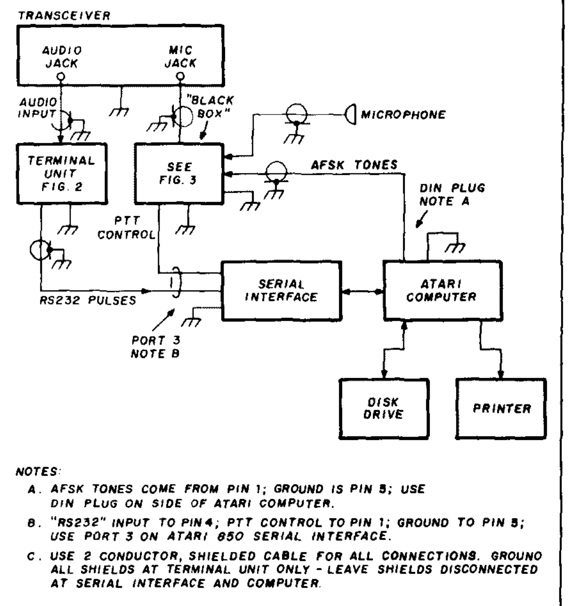


fig. 1. Cabling arrangement between the Atari computer, transceiver, and terminal unit.

tors and capacitors connected to this chip are used to change its frequency response characteristics. This circuit provides digital pulse strings when either low tone pairs or high tone pairs with a frequency difference of 170, 425, or 850 Hz are received.

All parts except the XR2211 chip come from Radio Shack; the XR2211 is available from Jameco.* Application note AN-01 from Exar Integrated Systems† explains chip operation. The serial interface used in this application is the Atari™ 850. With this interface, it is possible, under program control, to receive Baudot or ASCII at rates from 60 to 960 WPM. Although this interface is billed as an RS232-level device, it works fine with the 0 to 12 volt signal generated by the terminal unit described.

To transmit RTTY, there must be some way of choosing the character or number you wish to send. This is normally done via a keyboard. Pressing a keyboard button closes a switch which is detected by the computer program and decoded into a unique binary number. This number is normally converted into a pulse string, which is subsequently converted to either mark or space tones, depending on the voltage level of the pulses. These *audio frequency tones* must be held for the appropriate time (approximately 22 milliseconds for 60 WPM) and fed to the microphone input circuit of the transmitter.

detailed solutions

The audio tones sent to the microphone jack need to be fairly precise in frequency and duration. In this application, advantage may be taken of a feature in the Atari computer itself. The Atari has a set of internally programmable sound generators (they are used to simulate explosions, battle tanks, and so forth in game programs). These sound chips happen to generate the audio frequencies for mark and space at all frequencies and shifts needed. Although these tones are neither precisely those specified for RTTY (plus/minus 10 Hz) nor perfectly sinusoidal, they work flawlessly.

This feature makes it unnecessary to build an external tone generator — thus the receive-only terminal unit. To control the time duration of the tones a small assembly-language program was used. The BASIC language that composes most of the program is not fast enough to turn the tones on and off at the required speed.

The same assembly language program is used for all of the different tone duration times. The main BASIC *program modifies the timing* constants in this assembly-language program whenever you change from one WPM rate to another.

FCC regulations require that Amateur Radio operators provide CW identification at the end of their RTTY transmissions. This is accomplished using the same method as the tone pair generation. The program transmits the CW ID at approximately 20 WPM at a single pure tone that is between the mark and

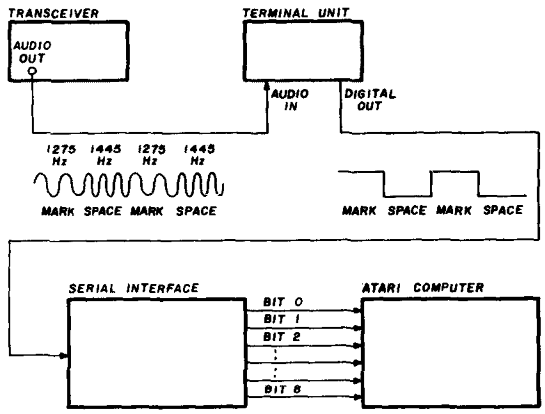


fig. 2. The terminal unit converts tones to a digital pulse train.

*Jameco Electronics, 1355 Shoreway Road, Belmont, California 94002.
†Exar Integrated Systems, Inc., 750 Palomar Avenue, P.O. Box 62229, Sunnyvale, California 94088.

space frequencies. This enables the receiving station to hear your CW ID without retuning the receiver.

Baudot computer program options

The program allows you to select any of several options, which include receiving RTTY; transmitting at 60 WPM using the low tone frequency pair, 170-Hz shift; transmitting at 60 WPM using the high tones frequency pair, 170-Hz shift (for VHF); transmitting at 100 WPM-low tones, 170-Hz shift; transmitting at 100 WPM-high tones, 170-Hz shift; printing using a hardcopy device; and "go to ASCII program."

Other options included in the program are:

A. Automatic transmitter turn-on/turn-off using the PTT feature, accomplished by using a spare pin on the Atari 850 serial interface. One of the signals available at the output of this interface is called Data Terminal Ready. This pin supplies either +12 or −12 volts and is switchable under program control. It is therefore ideal for driving a transistor switch to activate PTT when transmitting and deactivate the PTT when receiving (see **fig. 4**).

B. Brag tapes (pre-recorded messages) are nice to have, so there is a feature in the program that allows you to call up and transmit any one or all of seven Brag tapes stored on the disk. When you are transmitting, a Control A will read Brag tape 1 from the disk and send it. Control B sets Brag tape 2, and so on up to Control G. Control H is reserved for the CW ID To Follow announcement and automatically sends your CW ID. A separate program is used to build the Brag tapes.

C. Hard copy on a printer is possible. The program stores all received characters in memory and after the QSO allows you to list it to the printer. This application is programmed to store 4000 characters. It can be increased or decreased depending on memory availability.

D. Some systems permit transmission of date and time. Control T will do this if you enter the correct time and date into the program when it first runs. (This piece of coding is not smart enough to change the month if you transmit past midnight on the last day of the month — a good thing for you to modify!)

E. Sometimes a reception error occurs and you go into the Numbers printing mode erroneously. Pressing Start forces you back to Letters mode immediately.

F. Pressing Select clears the screen and printer storage buffer, and reprograms the serial interface to change the expected reception baud rate (WPM). You can cycle the WPM reception rate from 60 to 66 to 75 to 100 back to 60 with four pressings of the Select key. This is handy for copying commercial RTTY broadcasts.

G. The Option key aborts the receive portion of the program and allows you to begin transmission, begin printing, select a different WPM transmit rate, or go back to receiving.

H. Control I aborts the transmit section of the program and goes to the Option section without sending a CW ID.

I. Control H sends CW ID To Follow-DE (Your Call), in RTTY, then sends your call in CW and immediately

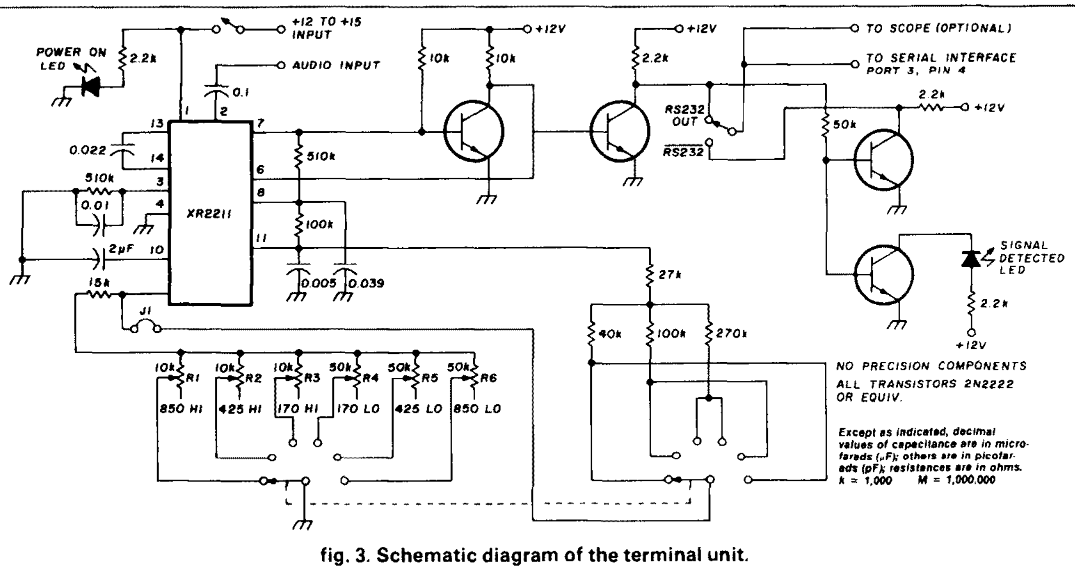


fig. 3. Schematic diagram of the terminal unit.

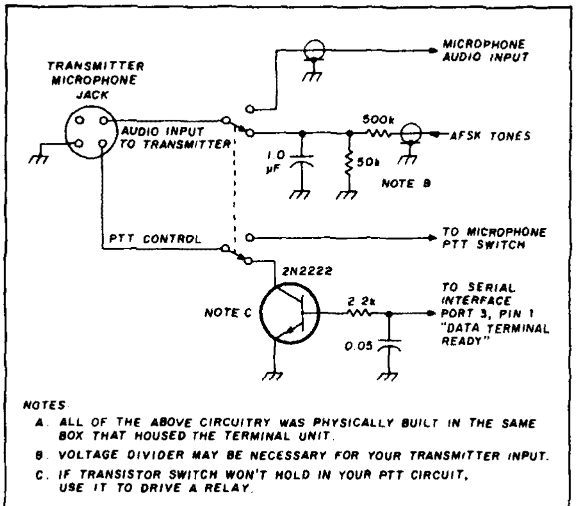


fig. 4. A method of activating PTT by means of the Atari computer serial interface.

switches to receive at the same baud rate you were using in transmission.

differences in the ASCII program

The ASCII program is similar to the Baudot program just described. Its option section permits: receiving ASCII; transmitting at 110 Baud, 170-Hz shift, low tones; transmitting at 300 Baud, 425-Hz shift, high tones (for VHF); transmitting at 600 Baud, 425-Hz shift, high tones; transmitting at 1200 Baud, 850-Hz shift, high tones; printing to a hardcopy device; "Go to Baudot" program. (Receiving and transmitting at Baud rates above 300 have not been tested extensively to date.)

All of the options described above except Control T and the Letters-mode-forcing exist in the ASCII program. All of the equipment remains the same as for the Baudot program.

future possibilities

Some additional attractions you may want to add to the program could be: split-screen viewing of both typing and reception simultaneously; buffering your input so it's not sent immediately upon entry, but in fast strings to impress your contact with how smoothly and fast you type; automatic logging to the disk of time, date, call, and other QSO information; and CW reception — hint: This could be done through the joystick input port using the terminal unit described and an assembly language program.

getting started

Copies of the three BASIC program listings and the assembly language program listing described above are available from *ham radio*.*

For those of you who don't want to type all of these program listings into your computers, I'll be happy to send them to you on a 5¼ inch floppy diskette. I'll customize your diskette with your name and call. (Sorry, I can't send cassettes — just disks.)†

terminal unit construction adjustments

The circuit was built on perfboard and wire wrapped. *No printed circuit board is available.* Layout is not critical. I would advise using a metal box enclosure and shielded cable. Open Jumper J1 as shown in **fig. 2**.

Use an ohmmeter to measure the resistance from pin 12 of the XR2211 chip to *ground*. As you change the six-position switch's location, adjust R1 to R6 to give the following ohm readings:

ohm reading	adjust	switch position
17825	R1	850-Hz-shift high tones
19445	R2	425-Hz-shift high tones
20568	R3	170-Hz-shift high tones
26738	R4	170-Hz-shift low tones
30558	R5	425-Hz-shift low tones
33422	R6	850-Hz-shift low tones

Replace J1. The application notes from Exar give a more elaborate tune-up method, but mine worked fine with the above procedure. My settings were ±2 percent of the above values. These resistances are theoretically calculated from Exar's design information.

conclusion

The programs and equipment described in this article have been in use since November of 1982. They have resulted in numerous RTTY QSOs on both the hf and VHF bands. If you have an Atari computer, try it on RTTY! Please feel free to write if you have questions or run into problems with the programs. Include an SASE; I'll do what I can to help.

acknowledgments

My thanks to N5IB, Jim Giammanco, who put me onto the XR2211 chip and to my daughters, Wendy and Melanie, who let me onto their computer long enough to develop these programs.

*For copies of the program listings, send a stamped (37¢), self-addressed #10 envelope to PROGRAM LISTINGS, *ham radio* magazine, Greenville, N. H. 03048.

†To order a program diskette, send $10.88 directly to the author, David W. King, K5VUV, 743 Rodney Drive, Baton Rouge, LA 70808. The price includes diskette, postage, and service fee.

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

Now that we have temporary operating privileges for the 10 MHz band, we can look forward to the opening of the 18 and 24 MHz Amateur bands, another outcome of the 1979 World Administrative Radio Conference; the bands encompass 18.068 to 18.169 MHz and 24.890 to 24.990 MHz. As this column is being written (late February), it looks as if these bands are far away indeed for U.S. Amateurs, unless somebody pulls a rabbit out of the hat.

Operation on the new bands is authorized in many European and South American countries, although to date activity has been sparse except on weekends. Most stations congregate around 18.07 MHz and 24.9 MHz. In California, European signals came through very well on both bands in the morning hours during the winter.

The Federal Communications Commission has adopted Docket 80-739 NPRM of December 30, 1982, and the planned action (for "action" you may read "inaction") includes use of these frequencies by the fixed services until July 1, 1989. There is no indication of any plan for implementation of the WARC Resolution 640, and no indication that any interim action is contemplated.

So here we sit, as the sunspot count slowly sinks toward the next minimum, due to arrive in a few years. If the FCC follows its present policy of inaction, by the time the bands are opened for Amateur Radio *they will be useless* for long-distance communications. The next sunspot minimum is predicted to cover the period 1985 through 1990, so if we do achieve operating privileges in these bands in 1989, they will be of little use to us until about 1992. That's nine years from now! If the FCC really wishes to aid Amateur Radio, they should amend Part 97 of the Rules to permit operation on these bands on a noninterfering basis now.

As far as Amateur interference to existing fixed stations is concerned, both bands are a wilderness. Despite the FCC count of stations authorized to operate on these frequencies, few do. Six months of listening has logged very few fixed-service stations, far fewer in fact than the number noted on the 10-MHz band before it was authorized for Amateur operation.

I hope this frustrating hang-up can be solved, if possible before the end of this year.

the Kenwood R-600 communications receiver

This is not a product review but rather two ideas for improving this interesting receiver.

My general-coverage Collins 51J-4 receiver seems to grow more massive as the years roll by. It is an invaluable adjunct to my station, as it provides a-m/CW and SSB reception over the range of about 480 kHz to 30 MHz. With multiple mechanical filters, it serves in a pinch as a good Amateur receiver, backing up my regular ham-band-only receiver. I'd had my eye on the Kenwood R-600 receiver (which weighs less than 10 pounds!) for some time, and I finally bought one as a tentative substitute for the 51J-4, which, in its steel cabinet, is a real boat anchor.

I was really pleased with the Kenwood: excellent sensitivity, readout to 1 kHz, and excellent audio quality for listening to shortwave broadcast, regular broadcast, or long-wave reception of local aircraft weather reports. The little receiver exhibited two characteristics, however, that I found improvements for.

First, when I used a random-length wire antenna, cross-talk and birdie problems were evident in the broadcast and the long-wave bands. I found that a 70-pF variable compression mica capacitor placed at the antenna terminal, in series with the wire antenna, proved to be the cure. The capacitor is simply adjusted for minimum cross-talk; it does not hinder shortwave reception at all.

Second, I noticed a peculiar buzz-

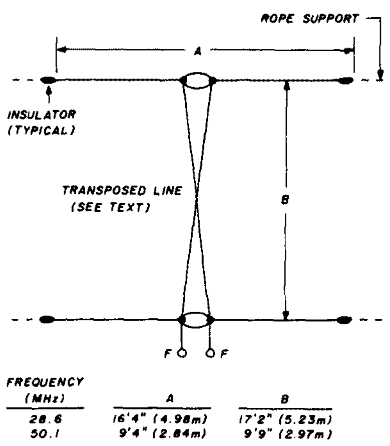


fig. 1. View of dipole stack beam for 10 or 6 meters. Directional pattern is into and out of page. Array is fed at F-F with 50 or 75 ohm coaxial line, as discussed in text. Coax connection at F-F should be waterproofed.

ing on the high-frequency bands, particularly around 20 MHz. The high-pitched buzzing noise grew loudest when I brought my hand near either the receiver's tuning dial or the digital frequency readout immediately above it.

It only took a moment to ascertain that the receiver was listening to the counting pulses that drove the digital frequency display. Moving the antenna about in the room alleviated the problem somewhat; and the use of an elevated dipole fed with coax a distance from the receiver completely eliminated the noise. But the dipole is useless for general coverage reception. What to do?

Using a short test lead as a probe connected to the antenna input terminal of the R-600, I found that the counter noise was coming from the glass dial of the frequency readout. Removing the top and bottom covers of the receiver enabled me to see that the readout was well shielded from the rear; but the shield was open to the front to make the readout visible.

My cure was quick, inexpensive, and simple. I removed the knobs and front panel (the panel is held in place by top and bottom bolts, plus two bolts under the tuning dial). At the hardware store I bought an envelope of "screen door patches," which are little squares of aluminum screening. I cut one of these squares down so that it was about 2 inches long and 1 inch high, just big enough to place behind the glass window. When the glass was replaced, it pressed the screen against the metal chassis, making a good ground connection. Before reassembly I sprayed both sides of the screen with flat black enamel to remove any reflection, leaving the edges of the screen clear of paint to make a good ground.

That did the trick! It bottled up the counter noise so well that it cannot be heard on any band.

Most modern ham equipment has some kind of frequency display. Does yours generate noise that can get into the front-end of the receiver? Perhaps some of those funny noises you've noticed from time to time are caused by this problem. You can make a quick check by disconnecting your regular station antenna and using a short pickup wire as a substitute antenna. Place the free end near the digital display and check it on all ham bands. If you hear any high pitched birdies, reconnect your station antenna and see if you can still hear them. If not, you probably have nothing to worry about. But if you do notice any counter noise, try a small piece of screening to bottle it up — provided the manufacturer shielded the readout assembly on the *inside* of your receiver.

wire antennas for 10 and 6

It's fun to build antennas! And you don't need an advanced degree in computer engineering to do it. There are plenty of simple wire antennas that you can build in a few hours, antennas that will outperform the popular ground plane or dipole. This is especially true on 6 and 10 meters, where high-gain antennas become a manageable size.

Shown in this section are two wire beam antennas for these bands. The first is a stack of dipoles and the second is a simple V-beam. Both designs were popular years ago but have been obscured by the rotary Yagi and quad.

Even if you don't have room or money for a rotary, you can build one of these simple beams for just a few dollars. They have a bi-directional (figure 8) pattern, like the dipole, and they provide worthwhile gain on both transmit and receive.

The dipole stack beam is shown in **fig. 1**. The array consists of two dipoles, one above the other, the lower dipole fed from a coaxial transmission line. The dipoles are cross-connected by an open wire line, as shown in the illustration. Power gain is about 4 dB or more over a dipole when the bottom of the antenna is at least one-half wavelength above ground. Dimensions for the two bands are given in the illustration. The two-wire interconnecting line is made of No. 16

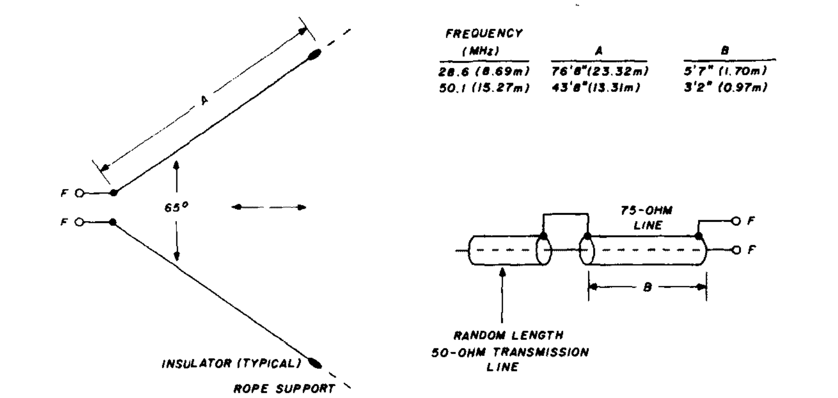


fig. 2. Plan view of the V-beam. Array is fed with 50-ohm line and 75-ohm transformer. Joint between lines and coax connection at F-F should be waterproofed.

enamel wires, spaced 3 inches apart. The spacers are made of Lucite® or plastic rods about 4 inches long. They are drilled to pass the wires, which are tied to the insulators with short sections of scrap wire. A half-twist is given to the line to reverse the connections at the opposite ends.

The stack beam is fed at points F-F with either a 50 or 75 ohm coaxial line. Feedpoint impedance at resonance is about 60 ohms so the SWR at antenna resonance should be well below 1.5-to-1 using either cable. The antenna is hung in the vertical plane, broadside to the direction or radiation. The coaxial line is wrapped into a four-turn coil directly below the feedpoint, to decouple the outside of the line from antenna currents. Keep the decoupling coil at right angles to the antenna wires.

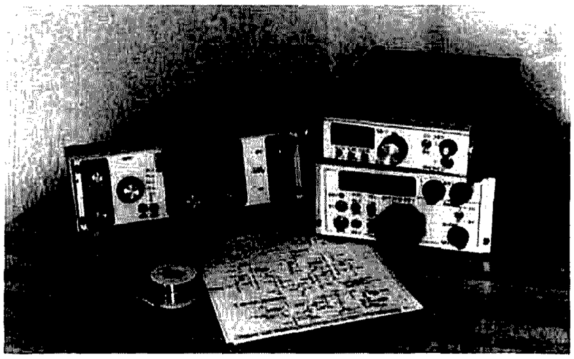
The bottom of the antenna should be at least as high above ground as dimension B — the higher the better.

The V-beam is shown in **fig. 2**. The wires are parallel to the ground and their length (2-1/4 wavelengths) plus the selection of the included angle between the wires provides a bidirectional array which shows a power gain over a dipole of about 4.5 dB. Feedpoint resistance of the antenna is matched by the use of a 50-ohm transmission line and a 75-ohm quarter-wave impedance-transforming section, as shown in the illustration.

The beam is constructed of No. 16 enamel wire. Either hard-drawn wire or prestretched softdrawn wire is recommended. The coaxial transformer section of the line is wrapped into a four-turn coil directly at the feedpoint to decouple the outside of the line from antenna currents. At the design frequency, the measured SWR on the line should be below 1.5-to-1. For best results the V-beam should be mounted at least one-half wavelength above ground.

One nice fact about both of these beam antennas is that they are virtually invisible once they are up in the air. That's a plus if you live in a neighborhood that has an anti-ham bias!

ham radio



Three different versions of the receiver, two of which have been expanded into transceivers. The one on the left is the original which was built in modules. It uses a cabinet available from Radio Shack, and a homemade front panel. On the top right is the basic receiver described in this article. It uses an inverted chassis with cover plate. Wooden rails have been added to both sides and an aluminum trim strip adds a finishing touch to the front panel. A bar graph display has been used here instead of an S-meter. It is mounted just above the digital readout. On the bottom right is a unit built by Bob Kirby, WA3DYF. His version includes an antenna tuner, so that a random length wire can be used as an antenna.

modular two-band receiver

State-of-the-art circuitry with digital frequency readout

I have often been impressed by the many excellent articles which have been published about my favorite subject — communications receivers. A problem I have found with most of the articles, however, is that duplicating some of the circuits is often difficult. Some receivers use surplus or discontinued parts, or parts not readily available. In some cases extremely expensive, custom-made components are used.

There is no reason why a top-quality, high-performance receiver should cost a small fortune to build — or require a bench full of sophisticated test equipment to adjust. You *can* build a receiver for less money than you would have to spend to purchase one of similar performance.

This article describes my answer to these problems. Here is a reliable, high-performance Amateur communications receiver that will perform as well as some of the best receivers available to Amateurs today. The basic two-band design can be expanded to cover the other bands, and, with the addition of two boards, can operate as a transceiver on CW and SSB.

the evolution of the design

The typical receiver should be able to handle strong signals (both on and off frequency), such as

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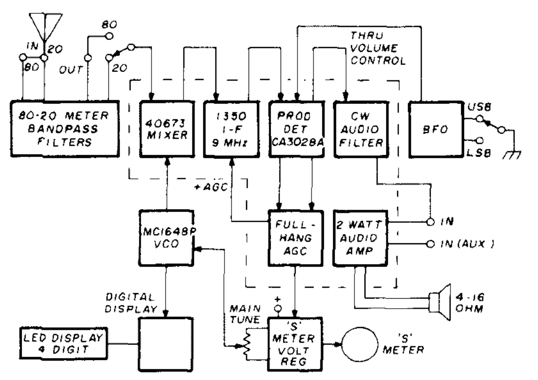


fig. 1. Block diagram of the SSB/CW receiver.

those typically found under Field Day type operation. Many otherwise excellent receivers sold to Amateurs are terrible at this. Manufacturers, in an attempt to improve poor dynamic range, add an attenuator to the front end of their receiver. This does not really cure the problem, but substitutes one problem for another. Yes, the receiver no longer overloads on strong signals, but now the operator can hear only the strongest signals. It should not be necessary to use an attenuator or to run an rf-gain control at less than maximum sensitivity under any type of operation encountered on the Amateur bands today.

Most interference heard on the lower Amateur bands is generated in the receiver. Poor dynamic range, as well as excessive gain in the front end, are usually to blame. Selectivity in the front end and i-f are also factors which are not given proper consideration in many designs.

A typical receiver should have excellent calibration. Digital readout is by far the best way to achieve this. Also, a digital-readout system eliminates the need for a mechanical dial drive, which can be extremely difficult to construct in the typical home workshop.

The receiver must have selectivity suitable for both SSB and CW reception. In this design, a KVG XF9B eight-pole crystal filter has been used. This filter is readily available, reasonably priced, and has excellent skirt selectivity, (see **table 1**). An audio filter with a bandwidth of less than 200 Hz can be switched in before the audio amplifier to help slice through the thickest QRM.

Any modern rig should, ideally, be able to operate on 12 volts dc. This not only simplifies portable operation, but during an emergency, this feature may save the day. Even during the worst disaster, a truck, auto, or motorcycle battery can provide enough power to get the communications started. The receiver draws only about 100 mA at normal volume level.

With emergency and portable operation in mind, small size and minimum weight are nice features to consider. One weight- and time-saving method involves eliminating the mechanical dial drive, readout, and tuning capacitor. I used a Jackson Brothers 6:1 reduction drive, which turns a ten-turn potentiometer, giving sixty turns to cover a 500-kHz band. The regulated voltage from this control is used to tune the VFO. Since the varactor diodes in the VFO require only a dc voltage, the packaging of the various boards needs not be influenced by any mechanical considerations. This packaging flexibility opens up a few new possibilities.

If mobile operation is contemplated, a remote-mount type of packaging could be used. The main receiver board, along with the VFO and BFO, could be in one box. The digital readout, tuning control, volume control, and S-meter, in a small box mounted under the dash, would complete the receiver. This idea is especially attractive for use in the small cars which are becoming so popular.

Finally, and of major importance, any circuit used in a receiver should be entirely reliable. By this I mean that only readily available, well-proven, solid-state devices should be used in circuits which are easy to duplicate without problems of instability. No changes or critical adjustments should be required to get the receiver working the first time.

In this design, I have relied heavily on the use of integrated circuits. This cuts size, complexity, and cost. The design has shown itself to be reliable and trouble-free. I know of no better way to put two pounds of circuitry into a one pound box.

the design

The two-band receiver consists of six printed circuit boards (see block diagram in **fig. 1**). The main receiver board contains the mixer, i-f filter and amplifier, product detector, AGC circuitry, active audio filter, and audio power amplifier. Other boards include the VFO, BFO, the voltage regulator and S-meter board, and the digital-readout board. Both of the bandpass filters are on one board.

table 1. Specifications for the KVG XF-9B filter.

application		SSB/RX
number of filter crystals		8
bandwidth (6 dB down)		2.4 kHz
passband ripple		<2 dB
insertion loss		<3.5 dB
input-output	R_t	500 ohms
termination	C_t	30 pF
shape factor		(6:60 dB) 1.8 (6:80 dB) 2.2
stop band attenuation		>100 dB

bandpass filters

Each band has its own double-tuned bandpass filter (**fig. 2**). This filter design has good rejection of unwanted signals both above and below the band of interest.[2] The two coils for each band are wound on ferrite cores and tuned with ceramic or mica trimmer capacitors.[3] (See **table 2.**) Once initially adjusted at the center of each band, the filters require no other tuning or adjustments.

One drawback of this type of front-end filter is the fact that the antenna must present a 50-ohm load.

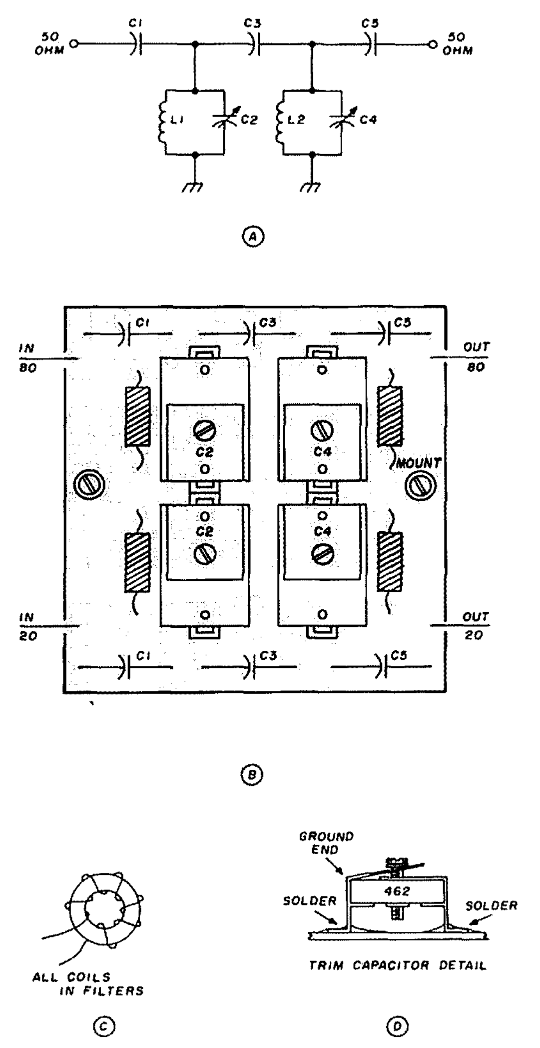


fig. 2. Schematic diagram (A) of the bandpass filters, and parts location (B). All coils are wound on a toroidal core as shown by (C); see table 2 for values. The tabs on trimmer capacitors are bent outward and soldered to the circuit board as in (D). All parts are on foil side of board.

Two more versions of the receiver, both of which have been expanded in function. The one on top has been designed for mobile use. It is built in a compact package measuring only 4 x 7 x 11 inches. The cabinet is formed by using two chassis fastened together with a top and bottom cover. A separate front panel hides the seam where the two chassis join together. The bottom unit is the dual-diversity unit mentioned in the article. Note the two dot displays to the left of the digital readout. These give a direct comparison of signal strength on each channel.

Severe mismatch at the antenna will detune the filter and cause a loss in sensitivity. It is not possible to just hang a wire on the antenna input and obtain good results. With a matched antenna, the filters are excellent.

table 2. Component values for the bandpass filters. C1, C3, and C5 are silver-mica capacitors. Coils are wound with No. 28 enamel wire.

	C1-C5	C3	C2-C4	L1-L2
80 meters	100 pF	12 pF	ARCO 464	35 turns on T37-2 core (red)
20 meters	15 pF	2 pF	ARCO 462	27 turns on T37-6 core (yellow)

the mixer

Initial experiments with the mixer stage involved double-balanced diode mixers, but these were rejected in favor of a dual-gate mosfet stage, as shown in the receiver-board schematic (**fig. 3**).

In theory, the diode double-balanced mixer is, perhaps, the ultimate design. However, in practice, the maximum capabilities of this device are rarely achieved in a home-built receiver.

The diode mixer, in order to work properly, must be terminated at all frequencies present — not just the i-f. This requires a circuit called a diplexer. This circuit can be very difficult to get working properly

with simple test equipment. This type of mixer exhibits a loss and also requires a high-level local-oscillator signal. This not only consumes extra power, but makes interstage coupling of the local oscillator signal a problem.

A dual-gate mosfet mixer, on the other hand, is not in the least bit temperamental, and good performance can be obtained without any adjustments. The drain is terminated in the eight-pole crystal filter. Impedance matching is handled by a 510-ohm resistor in the drain circuit of the mixer, which approximates the 500-ohm input impedance of the KVG filter.

the intermediate frequency amplifier

The local-oscillator signal and the desired incoming signal are mixed (heterodyned) to produce an output signal at the i-f center frequency of 9.0 MHz. This signal is then passed through the eight-pole crystal filter with a −6 dB bandwidth of 2.4 kHz. The outstanding skirt selectivity of this filter (1.8 shape factor) rejects off-frequency signals very well. It is this selectivity which allows you to separate the closely spaced signals which are common on the Amateur bands.

The signals at the output of the crystal filter must be amplified, of course, and this is handled by an integrated circuit which provides about 50 dB of gain and a bit more than 60 dB AGC control.

Although this eight-pin chip appears quite simple, the MC 1350 is really quite sophisticated.[1] It is also inexpensive. The gain of this stage is controlled by applying a voltage of 5 volts or greater to pin 5. An increase in voltage on this pin causes a decrease in gain in the chip.

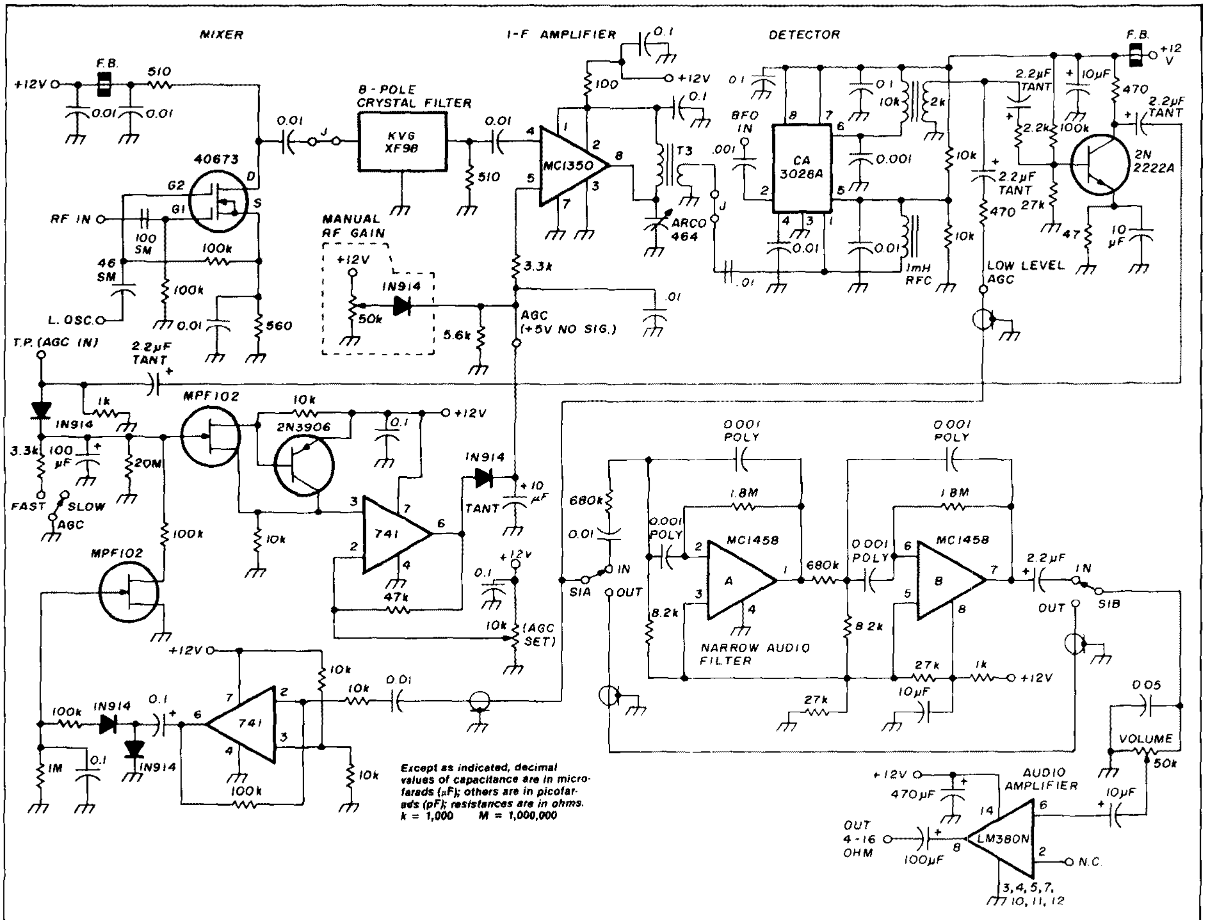


fig. 3. Schematic diagram of the receiver board of the modular receiver. S1 A and B is a miniature dpdt toggle switch. Capacitors in audio filter must be polystyrene or mica. Capacitors marked + can be either tantalum or other electrolytic. T3 primary is twenty-five turns No. 28 enameled wire on T37-6 core; secondary is five turns No. 28 wound over primary.

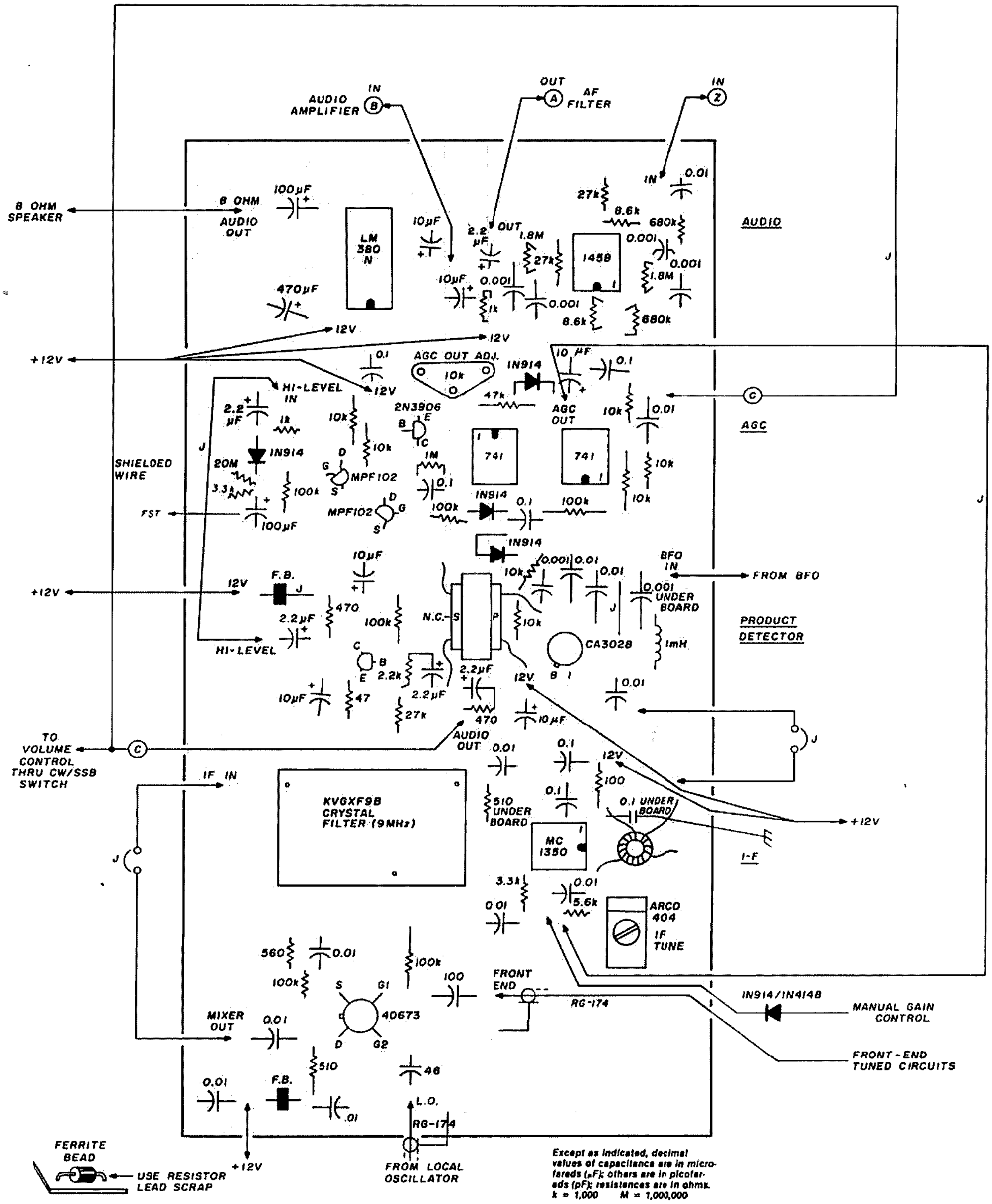


fig. 4. Component placement guide for the receiver board, viewed from the component side of the board with the printed circuit board in the background.

The printed circuit board (fig. 4) is designed so that the entire i-f amplifier stage, along with the crystal filter, can be diode or relay switched whenever this receiver is modified for use as a transceiver.

Although it was not included in the original design, a two-pole crystal filter was added at the output of the i-f amplifier. This was not necessary to realize excellent performance in the receiver, but it does produce a quieter receiver by eliminating most of the noise generated in the i-f amplifier. The use of the filter is especially noticed and appreciated when copying extremely weak signals near the noise floor of the receiver.

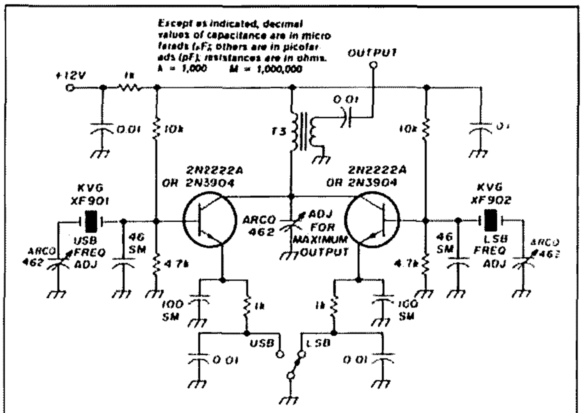


fig. 5. Schematic diagram of the BFO for the receiver. T3 primary is forty-five turns No. 28 enameled wire on T50-2 core; secondary is three turns No. 28 over primary.

The low cost involved by adding the two-pole filter is justified by the increased performance. The filter can be added without modification to the printed circuit board.

the product detector

I have experimented extensively over the past several years with direct conversion receivers (synchrodyne) and have found that the RCA CA3028-A integrated circuit works very well as a product detector. I have, therefore, used this device in this receiver. It exhibits good gain, low noise, excellent stability, low distortion, and a reasonable level of recovered audio. BFO level requirements are reasonable and non-critical. This chip also handles strong signals very well and this ability simplifies the design of the AGC system.

the audio stages

Detected signals from the product detector are coupled through an audio interstage transformer to the following stages. If more selectivity is desired for the reception of CW signals, the audio is routed through an audio filter.

Operational amplifiers have made filtering for selectivity at audio frequencies a practical method to use in the design of a new receiver or to improve an older receiver. This receiver uses a design based on an MC1458 dual-operational-amplifier integrated circuit. No critical parts are required, as experiments have shown that excellent performance can be obtained using typical 5 percent resistors and polystyrene capacitors. When it comes right down to it, it is

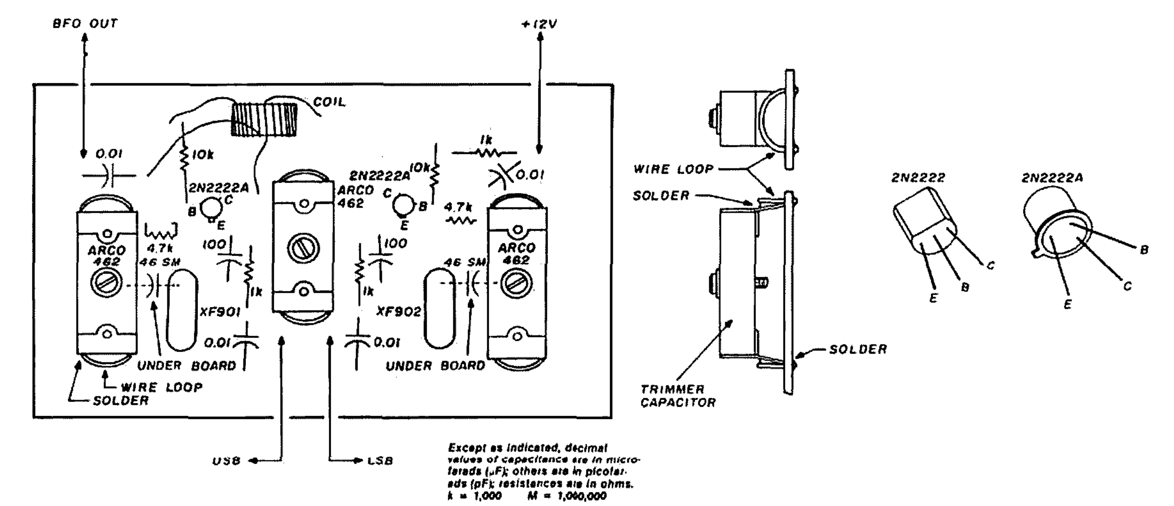
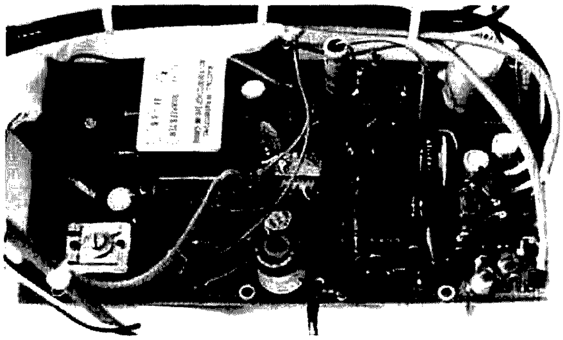
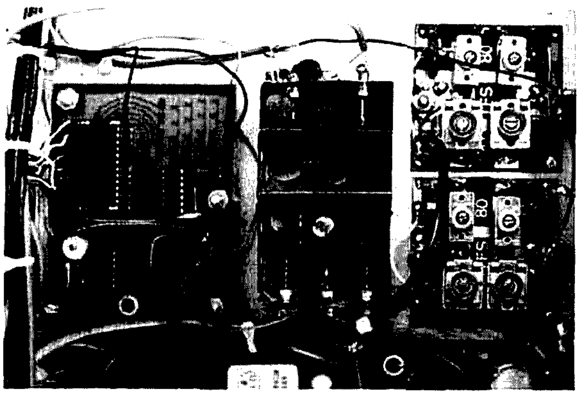


fig. 6. Component placement guide for the BFO board, shown from the component side with the printed circuit board in the background.



The main receiver board. Note the shielded wiring used on all audio and rf connections. From left to right are the mixer, crystal filter, i-f amplifier, product detector, AGC circuit, audio filter, and audio amplifier.



Three of the assembled boards. From left to right they are the digital readout board, the VFO, and the bandpass-filter board for 80 and 20 meters. A shielded control line goes to the VFO.

not important whether the center frequency is at 1.0 kHz or 1.1 kHz, or that the bandwidth at −6 dB is 200 Hz or 210 Hz. The design specifications call for a bandwidth of about 200 Hz at −6 dB, and a center frequency of 1.0 kHz. This is wide enough to eliminate any ringing tendency, yet narrow enough to cut through some of the worst interference.

One of my most basic concepts of receiver design is that simple is usually best. This idea is carried to the extreme when you consider the audio output stage. Only three parts are needed. The LM380-N integrated circuit will provide about 2 watts output in this configuration. It has low distortion, good gain, and is even thermally protected so you don't have to be concerned if the speaker becomes disconnected. I have used this receiver mobile and have found the audio output to be more than adequate when connected to an external speaker of good quality. The output stage will drive any load between 3 and 16 ohms. Don't ruin the excellent audio quality of this receiver by using an inferior speaker. Any of the many CB-type mobile speakers should be a good choice.

Incidentally, the audio chip has two input pins. One is used here, the other left floating. If the receiver is used as part of a transceiver, the other pin can be connected to the sidetone oscillator.

the AGC system

After weeks of experimenting with both audio- and rf-derived AGC systems, it became apparent that an audio-derived, full-hang AGC system worked best under signal conditions ranging from casual rag-chews to weak-signal CW work, DX pileups, and Field Day QRM.

Perhaps you have used a receiver and noticed that the S-meter (actually an indicator of AGC action in the receiver) would deflect up scale on signals not even detected in the audio output. This is typical of receivers using rf derived or i-f derived AGC systems that do not have sufficient selectivity ahead of the detectors for the AGC.

Because of this problem, the desired signal completely disappears or appears to become very weak because of the AGC action. Obviously, this is not an ideal situation. The receiver sensitivity should be totally controlled by the signal you wish to detect, not by QRM.

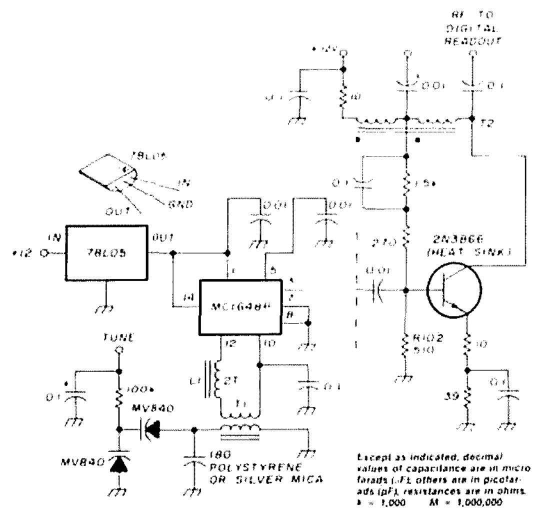


fig. 7. Schematic diagram of the VFO (local oscillator). T1 primary is thirty-five turns No. 28 enameled wire on T37-6 core; secondary is eight turns No. 28 over primary. T2 is ten turns No. 28 enameled wire wound bifilar on FT37-43 core. L1 is two turns No. 28 enameled wire on ferrite bead.

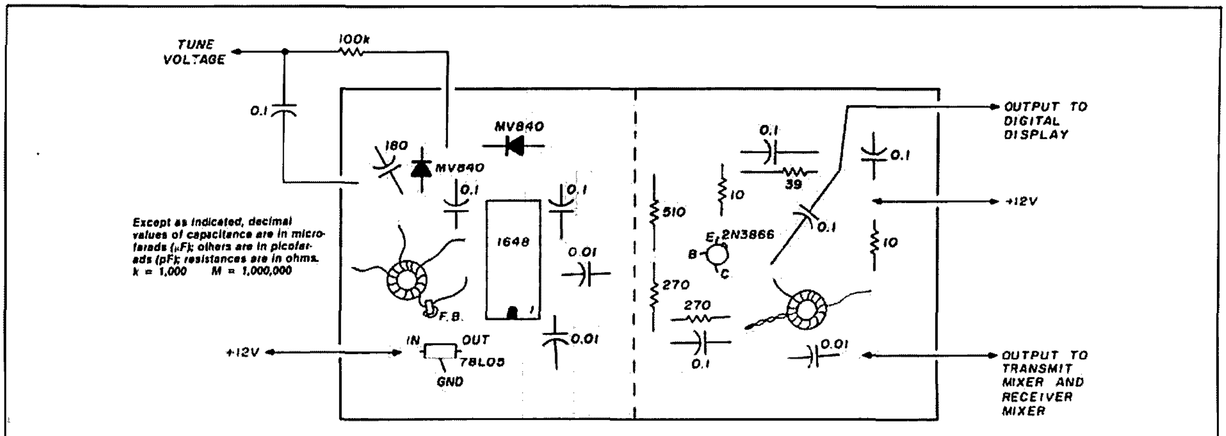


fig. 8. Parts placement guide for the VFO board with the printed circuit board in the background.

In practice, this is nearly impossible to do. But, through the use of selective filters and an audio-derived AGC system (as used in this receiver), this ideal comes closer to reality than you find in many commercial receivers.

A signal, first of all, must be detected and be present in the product-detector output to produce any AGC action. The strength of this signal determines how much AGC voltage will be applied to the i-f amplifier stage. When the need for a control voltage no longer exists, an FET switch is turned on, thereby shunting this voltage to ground, which brings the receiver back to maximum gain within a period of time determined by the time constants. The type or strength of the signals received does not affect this hold-in time. This type of circuit is discussed in greater detail in an ARRL publication.[1]

Two AGC time constants are available. The slower one is excellent for general SSB and CW use and the faster one allows good copy under adverse conditions.

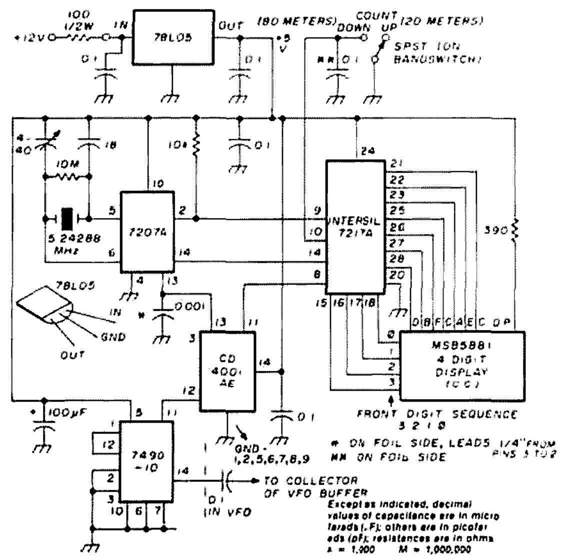


fig. 9. Schematic diagram of the digital frequency readout for the receiver.

the BFO

The beat-frequency oscillator is crystal controlled for stability. The circuit consists of two oscillators which share a common output tuned circuit (see **figs. 5** and **6**). The upper and lower sideband crystals are selected by grounding the appropriate control line. This board, like all the others, can be placed anywhere in the cabinet. Since only dc is being switched, it is not necessary to keep the control wires very short.

Each crystal has a trimmer capacitor so it can be set exactly on frequency. Another trimmer capacitor peaks the output tuned circuit at 9.0 MHz.

the VFO

Readers who are familiar with synthesized 2-meter equipment will probably recognize the MC1648 integrated circuit used in the VFO (**figs. 7** and **8**). It has become fairly common in VHF equipment but has not been used before, as far as I know, in a high-frequency receiver. It operates very well in this configuration.

One problem which may occur when using this integrated circuit is that it can oscillate above 250 MHz. The high-frequency oscillation is prevented by link coupling the tuned circuit to the IC through an rf choke. This low value inductance, as well as short lead length, proper pc board layout, and proper bypassing, prevents instability.

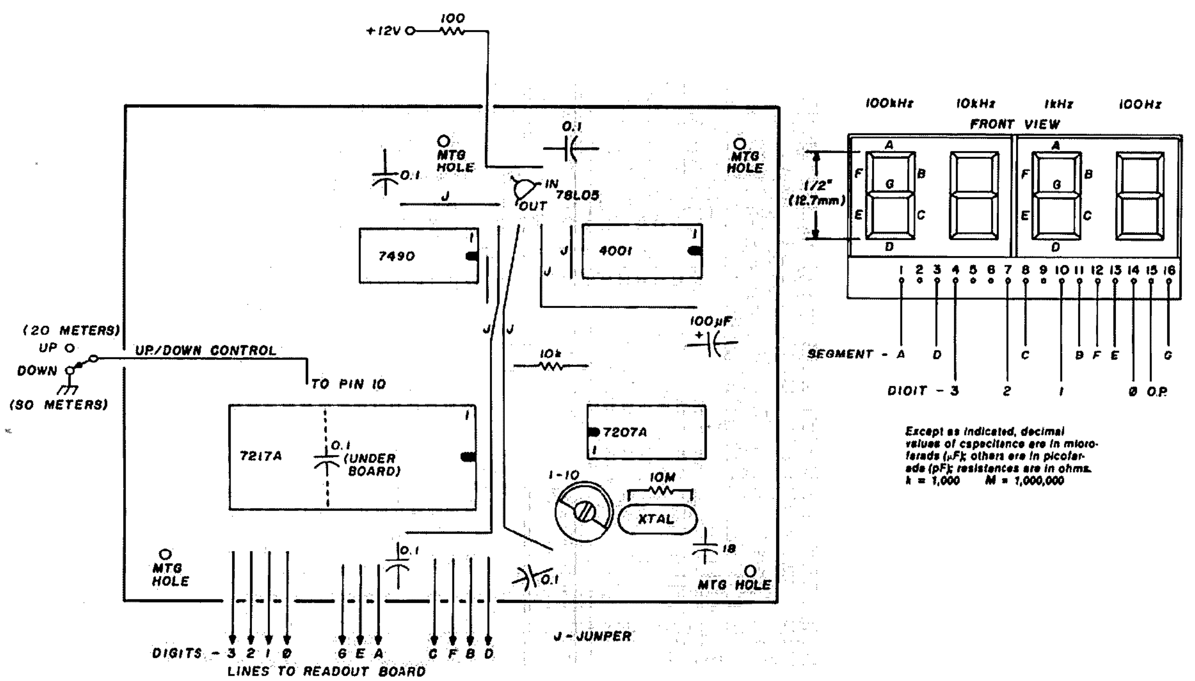


fig. 10. Parts placement of the frequency readout board with a printed circuit board in the background. The display is NSB5881 by National Semiconductor. To calibrate the readout, check against a freqency counter or calibrated receiver and adjust the 1-10 pF trim capacitor until readout is correct. Note that the first figure (MHz) of the VFO frequency (or the received frequency) is not displayed. Example: 5243.6 kHz (VFO) will be displayed as 243.6. Similarly, 14243.6 kHz will be displayed as 243.6 on the readout.

Two varactor diodes are biased by a regulated dc voltage which is controlled by a panel-mounted ten-turn potentiometer. A Jackson Brothers 6:1 vernier drive gives good bandspread.

VFO output is amplified and buffered by a Class-A 2N3866 stage. Output from the buffer is applied to the mixer stage in the receiver. Output to the digital readout is taken via a capacitor from the collector of this stage. This eliminates the need to add pulse shaping in the digital counter.

The regulated voltage, which is used to tune the VFO, is derived from a 6-volt, three-terminal integrated-circuit regulator which is mounted on the S-meter/voltage-regulator board.

Temperature compensation was found to be unnecessary for base-station applications. After a short warm-up period the drift is low enough to allow me to copy the ARRL RTTY bulletins without retuning. If you wish to use your receiver under adverse conditions, such as might be encountered during mobile operation, it may be necessary to add some sort of temperature compensation. Several schemes have been published and just about any of them will work. One simple method I suggest is to wire a 120-pF N750 ceramic capacitor in series with a low-value piston trimmer (approximately 2-5 pF) across the two varactor diodes. The trimmer should be adjusted to mid-range with a cold receiver. Hook a frequency counter to the VFO output and turn on the receiver. Plot the drift over about an hour's time. If drift is excessive, adjust the trimmer slightly, allow the receiver to cool and try the test again. This takes quite a bit of time, but once the magic combination is found, no further adjustment is needed.

digital readout

From the initial planning stages of this receiver, I decided to use a digital frequency display, but did not want the high current consumption, heat, or complexity of the usual designs. An Intersil LSI counter chip, along with three other integrated circuits, provides a four-digit readout with an accuracy of ± 100 Hz (**fig. 9**) with a components layout and printed circuit board shown in **fig. 10**.

The counter counts the VFO output and displays the last four digits. This corresponds to the frequency of the received signal. For example, a received frequency of 14,230.6 kHz is displayed as 230.6. On 80 meters, the counter counts down so that a received frequency of 3,976.8 kHz is displayed as 976.8.

Current consumption with this design is very low, and no noise from the counter can be heard in the audio output. The time base for the counter is a 5.24288 MHz crystal.

S-meter/voltage regulator

A meter amplifier designed to drive a low-current meter is included on this board (see **figs. 11** and **12**). The meters are readily available as CB surplus. Their current ranges are between 50 and 250 μA, and their cost is very low. A trim pot is used to set the meter to zero under no-signal conditions. Sensitivity of the amplifier is adjusted by changing the input resistor.

The board also holds a 6-volt, three-terminal integrated circuit and trim pots to set the upper and lower tuning range of the VFO. The trim pots should be set to allow a tuning range of about 4990 kHz to 5510 kHz. This range could be extended slightly to allow tuning in MARS or CAP frequencies.

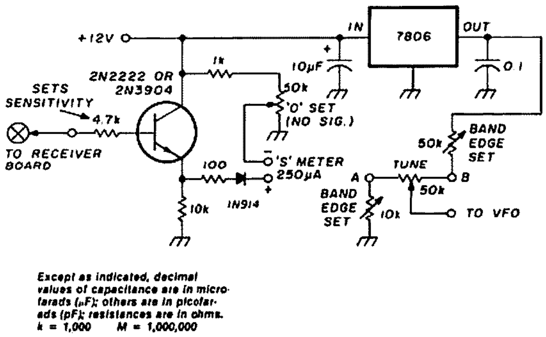


fig. 11. Schematic diagram of the voltage-regulator/S-meter board.

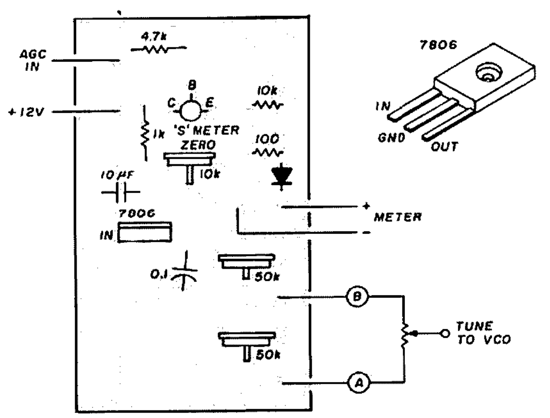


fig. 12. Parts placement guide for the voltage-regulator/S-meter board, component-side view with a printed circuit board in the background.

tuneup

Tuneup is a breeze!

1. Set BFO to frequency on either upper or lower sideband.

2. Peak i-f amplifier for maximum signal strength.

3. Peak bandpass filters for maximum at the center of each band.

4. Set AGC level at +5 volts with no signal on the input.

5. Set the timebase for the display on the digital display board so that the displayed frequency is accurate.

6. Set trimpots on the VR/S-meter board for the proper tuning range.

7. Set the zero adjust for the meter with no signal input.

8. Repeat as needed.

This receiver has been compared with some of the best available to Amateurs; in all cases it's held its own. The receiver sounds much quieter than any of the other receivers. Signals seem to pop out of the background. There is no roar of noise in the speaker when no signal is being received.

Single-tone dynamic range tests at 14.2 MHz work out to about 124 dB (**table 3**). This is with a signal

table 3. Specifications for the two-band receiver.

tuning range	3.5 to 4.0 MHz 14.0 to 14.5 MHz
VFO frequency	5.0 to 5.5 MHz (remotely tuned via dc)
i-f	9.0 MHz center frequency
BFO	USB: 8998.5 MHz LSB: 9001.5 MHz
digital readout time base	5.24288 MHz crystal
tuning resolution	± 100 Hz
voltage requirements	+ 12 Vdc; on-board regulation supplied as needed
current requirements	approximately 100 mA at medium volume setting
selectivity	SSB: 2.4 kHz (6 dB down) 1.8 shape factor (6.60 dB) 2.2 shape factor (6.80 dB) CW: peak type audio filtering; approximately 1-kHz center frequency with 6-dB bandwidth of 200 Hz
blocking	Better than 120 dB. (20-kHz spacing, 1-μV received signal strength at 14.2 MHz)

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More Details? CHECK—OFF Page 92

spacing of 20 kHz. A CW signal of 0.2 μV is very easily copied.

At the time of this writing, the receiver which I have described has been duplicated several times with consistent results. The receiver design has since been expanded to include two other boards which give it transceive capability on SSB and CW. This combination has been used to work forty-six states and several countries. Output power is four watts.

I have also designed a heterodyne-oscillator board that allows the receiver to be used on 160 through 10 meters.

packaging

The photographs show a few ideas for packaging your receiver. One uses a cabinet available from Radio Shack and other similar stores. Other versions are built into aluminum chassis which are used as cabinets. Surplus cabinets salvaged from old test equipment can be found for a minimal price. One of the receivers shown makes use of two standard Bud chassis (AC402 – 7 × 5 × 2 inches) assembled top-to-top with a front panel.

A very economic approach is to strip out an old low-cost receiver or transmitter. This will provide you with not only the chassis and cabinet, but also all the hardware you may need. Because of the design of the receiver, you need not worry about the mechanical arrangement of the various controls, as everything is switched with voltages. An old CB transceiver is another possibility. A new paint job and some rub-on letters will give a modern appearance. The only limit to the project is your imagination.

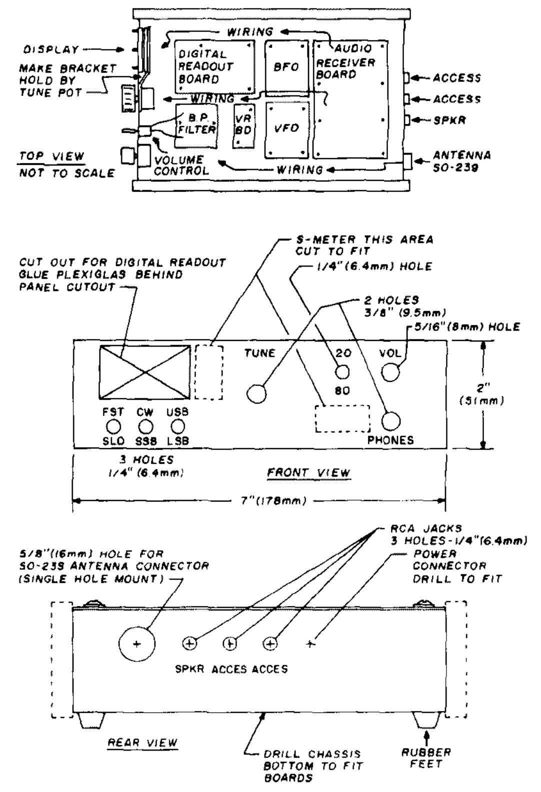

fig. 13. Suggested packaging for the two-band receiver. The enclosure is a standard 2 × 7 × 11 inch aluminum chassis (Bud AC402) with cover plate. Wood rails are added to the side and stained. The cabinet and panel can be painted to choice. Miniature toggle switches are used, and the volume control is a miniature type with on/off switch.

conclusion

Experiments have been performed using this design in a dual-diversity configuration, with excellent results. Basically, the design consists of one VFO board, one BFO, a digital frequency readout, two receiver boards, one audio stage, and a logic board to complete the hook-up.

This entire project has been approached from the viewpoint of an Amateur Radio operator, rather than as an engineer. It is relatively inexpensive and provides maximum performance at minimum cost, compared to receivers of similar performance. The design is easy to build, adjust, and package. None of the circuits are unstable, nor do they require any tinkering to achieve best performance. Best of all, the very nature of the design project promotes experimenting in the fascinating field of communication receivers.

As I have done with several of my projects, I have assembled several kits of parts for this two-band receiver. The kit includes all six pc boards and all parts needed to assemble them. A four-digit, ½-inch display and a Jackson Brothers vernier drive are also included. Documentation includes schematics, parts lists and layouts, block diagrams, and instructions. Drilling templates are provided for the version using a 2 × 7 × 11 inch (5 × 17.8 × 27.9 cm) chassis as a cabinet, **fig. 13**. The builder must supply the hardware, wiring, and cabinet. The cost of the kit is $320 here in the United States. Please send an SASE to the author with any inquiry.

references

1. Wes Hayward and Doug DeMaw, *Solid State Design for Radio Amateurs*, ARRL Inc., Newington, Connecticut 06111, pages 89, 92-94, 111-114, 117-119.

2. Doug DeMaw, *ARRL Electronics Data Book*, ARRL Inc., Newington, Connecticut 06111, page 55.

3. *Toroid Core Data Sheets*, Amidon Associates, 12033 Otsego Street, No. Hollywood, California 91607.

ham radio

Coming Events

ACTIVITIES

"Places to go..."

COLORADO: The Ski Country Amateur Radio Club's second annual Swapfest, July 23, Colorado Mountain College, 1402 Blake Avenue, Glenwood Springs. Free admission. Tables \$5.00 each. Talk in on 07/67. For further information: Frank, WA0BBI, Box 280, El Jebel, Colorado 81628.

ILLINOIS: The Hamfesters Radio Club is having its 49th annual Hamfest and Picnic, Sunday, August 14, Santa Fe Park, 91st and Wolf Road, Willow Springs, southwest of Chicago. Exhibits for OMs and YLs. Famous Swappers Row. Tickets \$3.00 at gate; \$2.00 advance. For tickets send check or MO with SASE to Hamfesters, P.O. Box 42792, Chicago, IL 60642.

ILLINOIS: The annual Belvidere Hamfest, Sunday, July 31, Boone County Fairgrounds, Highway 76, Belvidere. Tickets \$2.00 advance; \$2.50 at gate. Tables \$2.00 each. Saturday night camping. Talk in on 52 simplex. For information: Bob Anderson, K9DCG, 910 Locust Street, Belvidere, Illinois 61008.

ILLINOIS: The DuPage Amateur Radio Club's Hamfest/Computerfest, Sunday, July 10, 9 AM to 4 PM, Downers Grove American Legion Post grounds. Tickets \$2.00 at gate only. Large outdoor flea market. Plenty of parking. Refreshments available. Talk in on 144.89/145.49. For information SASE to: W9DUP, P.O. Box 71, Clarendon Hills, IL 60514. (312) 971-1156.

ILLINOIS: The Quad-Co Amateur Radio Club's 26th annual Hamfest of the "Breakfast Club", July 16 and 17, Terry Park, just east of Palmyra. Saturday night dancing and movies. Bring your basket lunch. Games, contests, golf and fishing. Bring your swap gear. Talk in on 3973 kHz from noon Saturday to 11 AM Sunday. Camping facilities from Friday afternoon to Monday AM. Pre-registration by July 7, \$1.50, \$2.00 at gate. Write Hamfest, c/o Quad-Co ARC, 602-D East Walnut, Chatham, IL 62629.

ILLINOIS: The Fox River Radio League Hamfest, the oldest in Illinois, Sunday, August 21, Kane County Fairgrounds, St. Charles. Exhibits, contests, demos and part of the flea market indoors. Additional outdoor flea market area. Tickets \$2.00 advance, \$3.00 at gate. Overnight parking Saturday, August 20, for campers and motor homes advance only \$3.00. Talk in on 146.94 simplex or 147.21/82 (Aurora). Campers, exhibitors, flea market space: George R. Isely, WD9GIG, 736 Fellows Street, St. Charles, IL 60174. Advance tickets Business SASE to Gerald Frieders, W9ZGP, 1501 Molitor Road, Aurora, IL 60505.

INDIANA: The combined LaPorte-Michigan City Amateur Radio Clubs will sponsor their Summer Hamfest, Sunday, July 17, LaPorte County Fairgrounds, State Road 2, 8 AM to 2 PM. Donation \$3.00 at gate. Refreshments. Indoor tables 40¢/ft. by reservation to P.O. Box 30, LaPorte, IN 46350.

KENTUCKY: The Bluegrass Amateur Radio Society will sponsor the Central Kentucky ARRL Hamfest, Sunday, 8 AM to 5 PM, August 14, Scott County High School, Longlick Road and US 25, Georgetown. Tech forums, awards, exhibits. Free outdoor flea market space. Tickets \$3.50 advance, \$4.00 at gate. For information/tickets: Edward B. Bono, WA4ONE, P.O. Box 4411, Lexington, KY 40504.

LOUISIANA: The Central Louisiana Amateur Radio Club will sponsor a Hamfest, Saturday and Sunday, July 30 and 31, Bolton Avenue Community Center, Alexandria. Swap tables available. For information: KA5HCJ, Central Louisiana ARC, P.O. Box 68, Alexandria, LA 71309.

MARYLAND: BRATS, the Baltimore Radio Amateur Television Society's famous Maryland Hamfest, Sunday, July 31, Howard County Fairgrounds, West Friendship, 15 miles west of Baltimore. Fairgrounds available for setup Saturday, July 30 at 2 PM. Overnight RV facilities. Talk in on 147.03 (+600), 146.76 (-600), 146.52 and 29.54/64. For table reservations and information: Mayer Zimmerman, W3GXK (301) 655-7812.

MICHIGAN: The Hiawatha Amateur Radio Association is celebrating its Golden Anniversary by sponsoring the 35th annual Upper Peninsula Hamfest, July 30, 9 AM to 5 PM, Michigan National Guard Armory, Ishpeming. Registration \$1.00. Tables available at \$3.00 each. Talk in on 146.16.76. Come and help us celebrate! For information: George Lehitnen, W8IOC, 100 N. R2, Ishpeming, MI 49849. (906) 485-5038.

MICHIGAN: The Amateur Radio Public Service Association of Saint Joseph County will hold its 5th annual Swap and Shop, Sunday, July 31, Saint Joseph County Fairgrounds, Centreville. Doors open 8 AM. Tickets $2.00 advance, $3.00 at gate. Indoor tables $3.00. Trunk sales free. Saturday night camping available at $6.00. Talk in on 52. For information: Warren Harder, N8EOX, 14820 Broadway Rd., Three Rivers, MI 49093.

MISSOURI: The 5th annual North Missouri Hamfest, sponsored by the *NEMO ARC, Kirksville and the Tri-*County ARC, Moberly, Sunday, August 7 at the air-conditioned Moberly Municipal Auditorium, Moberly. Inside flea market with limited number of free tables. Doors open for flea market and distributors 8 AM. Hamfest 9 AM to 3 PM. Tickets $2.00 at door, $1.50 advance. Refreshments served all day. Coffee and donuts for early birds. For information/tickets: Sam Fischer, KAØILO, P.O. Box 341, Moberly, MO 65270. Talk in on 147.69/09.

NEW HAMPSHIRE: Fly-in to NH's 3rd largest electronic flea market, Saturday, July 16, Manchester Municipal Airport. Starts 9 AM. General admission $1.00. Sellers $5.00 with own tables. Refreshments available. Pre-registration to New Hampshire FM Association, 30 Meadowglen Drive, Manchester, NH 03103. Talk in on 146.52 FM. For information: Dick DesRosiers, W1KGZ, (603) 668-8880 or Doug Aiken, K1WPM, 30 Meadowglen Drive, Manchester, NH 03103. (603) 622-0831.

NEW JERSEY: SCARC '83, the Sussex County Amateur Radio Club's fifth annual Hamfest, Saturday, July 16, Sussex County Farm and Horse Show grounds, Plains Road, off US 206, Augusta. General registration $2.00. Outdoor flea market space $4.00 advance, $5.00 at gate. Indoor sellers $5.00 advance, $6.00 at gate. Talk in on 147.90/30 and 146.52. Free parking. For information or registration: Lloyd Buchholtz, WA2LHX, 10 Black Oak Drive, RD 1, Vernon, NJ 07462.

NEW YORK: The Mt. Beacon Amateur Radio Club's Hamfest, Saturday, July 23, Arlington Senior High School, Poughkeepsie/LaGrange. Doors open 8 AM. Tickets $2.00. XYL's and kids free. Tailgating space $3.00. Tables $4.00 (1 free table and admission). Talk in on 146.37/97 and 146.52. For information: Art Holmes, WA2TIF, 2 Straub Drive, Pleasant Valley, NY 12569. (914) 635-2614.

NORTH CAROLINA: The Cary Amateur Radio Club's 11th annual Mid-Summer Swapfest, Saturday, July 16, Lion's Club Shelter next to Cary Senior HS, Cary. 9 AM to 3 PM. Free admission. Buy, sell, trade. Open auction. Talk in on 146.28/.88; 147.75/.15; and 146.52. For information: Cary ARC, P.O. Box 53, Cary, NC 27511.

NORTH CAROLINA: The Western Carolina Amateur Radio Society's all new Hamfest and Computer Fair, July 30 and 31, Buncombe County Firemen's Training Center, Asheville. Open 9 AM. ARRL booth, seminar by Bob Grove, WA4PYO, CW competition, RV parking and free camping (no hookups). See the latest in computer hardware and software. Talk in on 31/91, 16/76, 52 simplex. For ticket information: Garland Lance, NC4N, 854 Sandhill Road, Asheville, NC 28806.

OHIO: The 19th annual Wood County Ham-A-Rama, Sunday, July 17, Wood County Fairgrounds, Bowling Green. Gates open 8 AM. Free admission and parking. Trunk sales. Refreshments available. Dealer tables $5.00 advance registration. Saturday setup until 8 PM. K8TIH talk in on .52. For information, dealer rentals, SASE to: Wood County ARC, c/o Craig Henderson, Box 366, Luckey, OH 43443.

OHIO: The Northern Ohio Amateur Radio Society's NOARSFEST, Saturday, July 23, Lorain County Fairgrounds, Wellington. 8 AM to 5 PM. Donations $2.50 advance, $3.50 at gate. Children under 12 free. Blacktopped flea market area, $1.00 per car space. Free general parking. Refreshments available. Free overnight camping Friday night, no hookups. Mobile check-in, K8KRG, 146.52/52. Directions and info 144.55/145.15. For *information/* tickets: NOARSFEST, P.O. Box 354, Lorain, OH 44052.

OREGON: The 8th annual Lane County Ham Fair, July 16 and 17, Oregon National Guard Armory, 2515 Centennial, Eugene. Doors open 8 AM each day. Tech seminars, swap tables, 2 meter Bunny Hunt, kids' activities, computer demos. All day snack bar. Free parking for RV, no hookups. Saturday pot luck supper. Talk in on 52-52, 146.28/88, 147.86/26. For tickets and tables: Tom Temby, WB7WPU, 3227 Crocker Rd., Eugene, OR 97404. (503) 689-1761. Checks payable to Lane County Ham Fair.

PENNSYLVANIA: Nittany Amateur Radio Club's Hamfest & Computer Faire, July 9, New Location, Pleasant Gap Firemen's Park, Route 144, Pleasant Gap. Gates open 8 AM. All day technical operating sessions. Large tailgating area. Tickets $3.00. Tailgaters $5.00. Talk in on 146.16/.76 and 146.25/.85. For information: Dave Buckwalter, KC3CL, 1635 Circleville Rd., State College, PA 16801. (814) 234-0759.

TEXAS: The Austin ARC and the Austin Repeater Organization will sponsor Summerfest '83, August 12, 13 and 14, Austin Marriott Hotel, I-35 at Highway 290. Exhibits, meetings, indoor swapfest. Outdoor family activities. Admission $5.00 advance; $8.00 at door. Swapfest tables available at door. Reserved swapfest tables $1.00 advance. Talk in on 146.34/.94. For information: Austin Summerfest '83, P.O. Box 13473, Austin, TX 78711.

WEST VIRGINIA: The Triple States Radio Amateur Club *will present its 5th annual Wheeling, WV Hamfest at* Wheeling Park on Sunday, July 24, from 9 AM to 4 PM. Dealers, flea market and auction, free parking, refreshments, ARRL, SWOT booths, etc. Admission $2.00, children under 12 free. Indoor display, tables available, price of admission only but reserve space. CONTACT: TSRAC, Box 240, RD 2, Adena, OH 43901. Phone (614) 546-3930.

WASHINGTON: The Western Washington DX Club, W7FR, hosts the 31st annual Northwest DX Convention, Friday, Saturday and Sunday, July 29, 30 and 31, Double Tree Plaza Hotel, near South Center Shopping Mall and Seattle-Tacoma Airport. Saturday night banquet, Sunday morning breakfast. Speakers, slides, symposia and more. For registration: Ruth Bennett, WA7RVA, 6729 Beach Drive S.W., Seattle, WA 98116. (206) 932-1335.

WYOMING: The 1983 ARRL Rocky Mountain Division Convention in conjunction with the 51st W.I.M.U. Hamfest, August 5, 6, and 7, Virginian Motel, Jackson. Talk in on 146.22/82 and 3923 kHz. For more information: R.L. "Pete" Stull, WB7AMP, (307) 382-9023 or Dave Gregory, N7COA, (307) 875-5324.

WYOMING: Fourth annual High Plains Ham Roundup, September 9 and 10, Medicine Bow National Forest, 10 miles east of Laramie, I-80. Enjoy a real Western Ham Roundup. Bring your own food and drink. Roast beef furnished for Saturday pot luck supper. Blue Grass band, barbershop quartet and sing-a-long. Talk in on 146.25/85, 146.22/82 or 146.52 simplex. For information: Mick Marchitelli, P.O. Box 731, Laramie, WY 82070.

MONTANA-ALBERTA: The 49th Glacier-Waterton International Hamfest, July 15-17, H.O. at Waterton Homestead Campground, north of Waterton National Park entrance, Alberta, Canada. Bunny hunt, tech sessions, entertainment, swap tables. For information/pre-registration: P.O. Box 148, Milk River, Alberta, T0K 1M0.

BRITISH COLUMBIA: The Okanagan International Hamfest, July 30 and 31, Oliver Centennial Park, Oliver. Activities from Saturday, 1 PM, to Sunday, 2 PM. Entertainment, bunny hunts, pot luck luncheon Sunday. Talk in on 34/94 OKN Repeater - 76/76. For information: John Juul-Andersen, VE7DTX, 8802 Lakeview Dr., Vernon, BC V1B 1W3 or Lola Harvey, VE7DKL, 584 Heather Rd., Penticton, BC V2A 1W8.

RADIO EXPO: Sponsored by the Chicago FM Club, Saturday and Sunday, September 24 and 25, Lake County Fairgrounds, Routes 120 and 45, Grayslake, Illinois. Flea market opens 6 AM. Exhibits open 9 AM. Indoor flea market tables available at $5.00 per day. Tickets $3.00 advance, $4.00 at gate, good for both days. Seminars, tech talks, ladies' programs. Talk in on 146.16/76, 146.52 and 222.5/224.10. For information: SASE to Radio Expo 83, Box 1532, Evanston, IL 60204 or (312) 582-6923.

OPERATING EVENTS

"Things to do..."

JULY 3 AND 4: The Hannibal ARC will issue a third annual special certificate from the National Tom Sawyer Days celebration in Mark Twain's boyhood home town, *Hannibal, Missouri.* Hours: 1500-2100 UTC both days. Frequencies: Phone 7.245, 14.290, 21.400, 28.770. CW 7.125 and 21.125 MHz. To receive the certificate send large SASE and personal QSL card confirming contact to Hannibal ARC, WØKEM, 2108 Orchard Avenue, Hannibal, MO 63401. For further information: Tony McUmber, 2108 Orchard Avenue, Hannibal, MO 63401. (314) 221-6199.

JULY 4 AND 5: High Plains ARC will operate K7YPT at the historic Fort Laramie from 0000Z July 4 to 0000Z July 5. Frequencies: Phone 3.900-3850, 7.250, 14.250-14.300, 21.300-21.380. Certificate for large SASE to: K7YPT, Rt. 2, Box 303, Torrington, WY 82240.

JULY 9: The Waterville, NY, Central School ARC, WD2ALL, will operate from 1300-2000 UTC to commemorate the birth of George Eastman of Photography fame. Frequencies: lower portion of General phone and Novice CW bands. FM operation also planned for 146.52. Certificate and Club QSL available for SASE to WD2ALL via Callbook.

JULY 9 AND 10: The Cascades ARS (CARS) in conjunction with the Michigan Space Center in Jackson, is offering a Space Day certificate to all stations who work WB8CSQ on 3.900, 7.235, 14.285, 21.360 and 28.510 starting 0000 GMT July 9 through 1700 GMT July 10. Mail log information and $1.00 contribution for postage and materials to: CARS, Space Day 83, P.O. Box 512, Jackson, MI 49204.

JULY 16 AND 17: Wapakoneta, Ohio. Reservoir ARA will operate K8QYL from 1400Z July 16 to 0400Z and from 1400-1900Z July 17 from the home town of astronaut Neil Armstrong. Frequencies: Phone 7.260 and 14.285 MHz ± QRM. Certificate for QSL and large SASE to: K8QYL, P.O. Box 268, Celina, Ohio 45822.

JULY 16 AND 17: The Eastern Michigan Amateur Radio Club, K8EPV, will commemorate the annual Port Huron to Mackinac Island Yacht Race. Operation begins 10 AM EST (1500Z) through 10 PM EST (0300Z) on Saturday and Sunday. Frequencies: 3910, 7235 and 14285 phone; 3710, 7110 and 21110 CW. For an attractive certificate send legal size SASE to: K8EPV, 654 Georgia, Marysville, MI 48040; or C.B.A.

JULY 23: The Miami County ARC of Peru, Indiana, will operate K9ZEV in celebration of the 24th annual Peru Circus City Festival. Operation primarily in the General class SSB portion of 40 meters from 1400 to 2300 UTC. Check on 20, 15 and 10 meters as conditions permit. For a special commemorative QSL card send SASE to: Les Cattin, KA9FMZ, 163 W. Third Street, Peru, IN 46970.

JULY 30: The Tuscarora Amateur Radio Association will operate KI3D from 1200Z to 2400Z, from the National Historic Site of Tuscarora Academy, established 1839. Frequencies: 10 kHz up from lower edge of the General phone bands. Certificate for business SASE to: William Bratton, Box 31E, Star Route, Mifflintown, PA 17059.

JULY 30: The Tank-Automotive Command ARC will operate W8JPW from 1300-2000Z to commemorate the 42nd year of the Detroit Arsenal, home of the nation's first defense plant and the US Army Tank-Automotive Command. Frequencies: Phone 7.250-7.274, 21.400 and 146.49 MHz. CW 7.055 from 1500-1700Z. Send 9 × 12 SASE for unfolded certificate to: W8JPW, US Army Communications Command, Att: CCNC-TAC-M, 28251 Van Dyke, Warren, MI 48090.

JULY 30 AND 31: The Pike County ARC will operate W9CZH from the Lincoln Boyhood Memorial, Lincoln City, Indiana, from 1700Z July 30 to 1700Z July 31. Frequencies: 3.925, 7.265, 14.305, 21.395 phone; 14.090 RTTY; 146.52 FM; 7.133 CW. A special QSL will be issued for your QSL and SASE to KC9VH, Box 311, RR 1, Winslow, IN 47598.

JULY 30: Reservoir ARA will operate KR8M from 1330-1900Z from the Courthouse steps during the Celina, Ohio, Lake Festival. Frequency: 7.260 ± QRM. Certificate for QSL and large SASE to: KR8M, P.O. Box 268, Celina, Ohio 45822.

AUGUST 6 AND 7: The 21st annual Illinois QSO Party sponsored by the Radio Amateur Megacycle Society (RAMS) from 1800Z August 6 to 2300Z August 7, rest period 0500Z to 1200Z August 7. Frequencies: CW — 40 kHz from low end. Phone — 3890, 7230, 14280, 21375 and 28675. Novice — 25 kHz from low end. Exchange RST and County by Illinois stations. RST and state, province or country by others. For filing and further information: RAMS, K9CJU, 3620 N. Oleander Avenue, Chicago, IL 60634.

AUGUST 13, 14 AND 15: The 24th annual New Jersey QSO Party sponsored by the Englewood ARA. From 2000 UTC Saturday August 13 to 0700 UTC Sunday August 14 and 1300 UTC Sunday August 14 to 0200 UTC Monday August 15. Phone and CW same contest. A station may be contacted once on each band — phone and CW are considered separate bands. No CW contacts in phone band segments. General call "CQ New Jersey" or "CQ NJ". Suggested frequencies: 1810, 3535, 3900, 7035, 7135, 7235, 14035, 14280, 21100, 21355, 28100, 28610, 50-50.5 and 144-146. For filing or information: Englewood Amateur Radio Association, P.O. Box 528, Englewood, NJ 07631.

AUGUST TO DECEMBER 1983: Jamaica Amateur Radio Association Award commemorating Jamaica's 21st year of independence, August 6, 1983. This award is available to all licensed Amateurs for CW, phone or mixed modes. Rules: Contact with 5 different 6Y5 stations, any band, August to December 1983. Submit QSL cards or written proof with time, date, band, mode and 6Y5 stations worked and fee of $3.00 U.S. or 10 IRC's and 8 × 10 SASE to: Awards Chairman, Gerald Burton, 6Y5AG, Box 214, Kingston 20, Jamaica W.I.

More Details? CHECK—OFF Page 92

audio filter building blocks

Active filters in theory and practice

In single-sideband and CW communications, the received audio signals are simple frequency-translated versions of the rf signal received at the antenna. This translation is accomplished by one or more mixer stages. The receiver block diagram usually includes an intermediate-frequency (i-f) stage that does most of the filtering to obtain selectivity. That is, this stage passes the desired signal on through but rejects any unwanted signals.

Fig. 1 is a block diagram of a simple receiver, which consists of a mixer and variable oscillator, i-f amplifier/filter, product detector, oscillator, and audio amplifier. The mixer and its variable oscillator translate the incoming signal from its original frequency to the i-f frequency. The i-f amplifier is also labeled as a filter since it has a bandpass frequency response and performs most of a receiver's filtering for selectivity. The output of the i-f stage is translated by the product detector to audio frequencies which are then fed to the audio amplifier and speaker. Since the signal present at the audio amplifier is a frequency-translated version of the signal at the i-f stage, filtering at the audio stage is equivalent to filtering at the i-f stage. Thus, receiver selectivity can be improved by adding an audio filter between the output of the receiver and the speaker or headphones.

In practice, audio filtering has a few disadvantages when compared with i-f filtering. Any automatic gain control (AGC) action that takes place in the i-f because of a strong interfering signal may wipe out the desired signal, regardless of how good the audio filtering may be. Also, any distortion introduced in the i-f system due to interfering signals cannot be completely eliminated by audio filtering. However, audio filtering does improve reception and, since it can be added externally, no receiver modifications are necessary.

building blocks

Here are some basic building blocks which can be used either individually or in cascade to produce a filter which meets your needs. These filters will all have unity gain (0 dB) in the passband to simplify their interconnection. All of the op-amps have been designed to use a single 12-volt supply. The circuits draw little current (typically 10-20 mA), so any simple power supply or battery can be used.

cw filter

A very simple active audio filter for CW can be made using a state-variable filter (see **fig. 2**).[1] This filter has a bandpass characteristic which can be of

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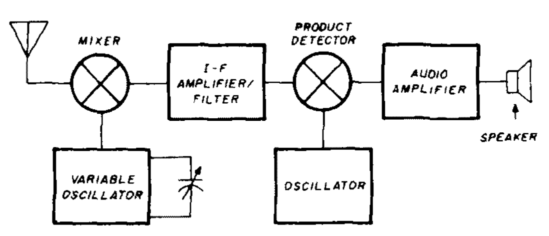


fig. 1. Block diagram of a simple receiver. Signals at the antenna are frequency-translated down to audio.

table 1. Equations for bandpass filter.

$$\frac{V_{out}}{V_{in}} = -\frac{1}{R_oC_2}\left(\frac{S}{S^2 + S\left(\frac{1}{R_QC_2}\right) + \frac{R_4}{R_1R_2R_3C_1C_2}}\right)$$

$$bandwidth\ (Hz) = \frac{1}{2\pi R_QC_2}$$

$$center\ frequency\ (Hz) = \frac{1}{2\pi}\sqrt{\frac{R_4}{R_1R_2R_3C_1C_2}}$$

$$passband\ gain = \frac{R_Q}{R_o}$$

fairly high Q (very selective), and the center frequency of the filter can be varied using one variable resistor. The bandpass can also be varied, but two resistance values must be changed to keep the bandpass gain constant. The values shown give a 3-dB bandwidth of 100 Hz and 400 Hz, although other bandwidths can be produced by changing R_0 and R_Q, which must remain equal to preserve unity gain. The design equations for the filter are given in **table 1**. Also, be aware that decreasing the bandwidth much beyond 100 Hz is likely to result in an oscillator instead of a filter because of the less-than-ideal nature of op-amps. The LF356 op-amp (which is a fairly wideband device) was used to minimize these effects. With a lesser op-amp, the filter will have a more peaked response at higher center frequencies and the bandwidth will not be constant as the center frequency is varied. As with all high-gain, wide-bandwidth devices, be sure to keep the power supply well bypassed (a 0.1-μF ceramic capacitor near each IC).

This particular configuration can be adapted to a notch filter by adding just one op-amp. This op-amp is configured as a summing amplifier which adds together the output of the bandpass filter and the input to the system. Since the bandpass-filter output is inverted (180-degrees phase shift) relative to the input, the net result is that the bandpass output is subtracted from the input. This results in a notch filter, since the signals in the passband of the bandpass filter cancel when the inverted and non-inverted signals combine.

The depth of this notch is limited by the matching of the gain-setting resistors in the summing amplifier and also in the bandpass filter. Therefore, the 10-kilohm variable resistor was included to allow some compensation for gain errors. The notch depth can be adjusted by tuning in a carrier or crystal calibrator on a receiver, adjusting the tune control to notch out the carrier, and then adjusting the 10-kilohm variable resistor for minimum audio signal. The minimum notch will probably not occur at the same setting for both bandwidths, but tuning with one bandwidth should result in an adequate notch on the other.

SSB filter

An audio filter for use with single sideband can be built using only two op-amps. One op-amp is configured as a highpass filter with cutoff frequency around 300 Hz, and the other is configured as a lowpass filter with a cutoff frequency of about 3 kHz. This results in a bandpass characteristic encompassing the standard audio frequency range for voice transmission.

The design equations are given so that other highpass and lowpass cutoff frequencies can be used. A Q of 1 was chosen so that the peaking in the passband is limited to about 10 percent. For simplicity, all capacitors are of equal value in the lowpass filter. The design equations for these filters are given in **table 2**. The op-amps in this case can be one like the LM307, since the gain-bandwidth demands of the circuit are not excessive.

These two filters can, of course, be used separately. The highpass would be useful for filtering out 60-Hz hum from an older tube-type rig, and the lowpass alone will help most any sideband rig in reducing the high-frequency adjacent-channel interference.

table 2. Equations for SSB filter.

Equations for highpass section.

$$\frac{V_{out}}{V_{in}} = \left(\frac{S^2}{S^2 + S\left(\frac{3}{R_2C}\right) + \frac{1}{R_1R_2C^2}}\right)$$

when $R2 = 10R_1$

and

$Q = 1$

$$f_{3dB} = \frac{0.77}{6\pi R_1C}$$

Equations for lowpass section.

$$\frac{V_{out}}{V_{in}} = -\frac{1}{R^2C_1C_2}\left(\frac{1}{S^2 + S\left(\frac{3}{RC_1}\right) + \frac{1}{R^2C_1C_2}}\right)$$

when $C_1 = 10C_2$

and

$Q = 1$

$$f_{3dB} = \frac{1.3}{6\pi RC_2}$$

driving headphones

All of these circuits can be used to drive headphones without an additional amplifier stage. **Fig. 4** shows a circuit to be used for connecting virtually any headphone to the output of an op-amp. The capacitor blocks the dc voltage that is present at the output of the op-amp, and the two resistors act as a voltage divider to reduce the level into the headphones. Most headphones are so sensitive that they need very little drive, so the signal is attenuated by these resistors.

driving speakers

Fig. 5 shows a simple audio amplifier which uses one-half an LM1877 stereo-amplifier IC. The output of any of the filter sections can be used to drive the input of this amplifier. This is one of many audio-amplifier ICs that are ideal for this sort of application. This circuit was taken directly from the manufacturer's data book[2] and care should be taken in adjusting any of the values since the device is not necessarily stable at unity gain. Care should also be taken in by-

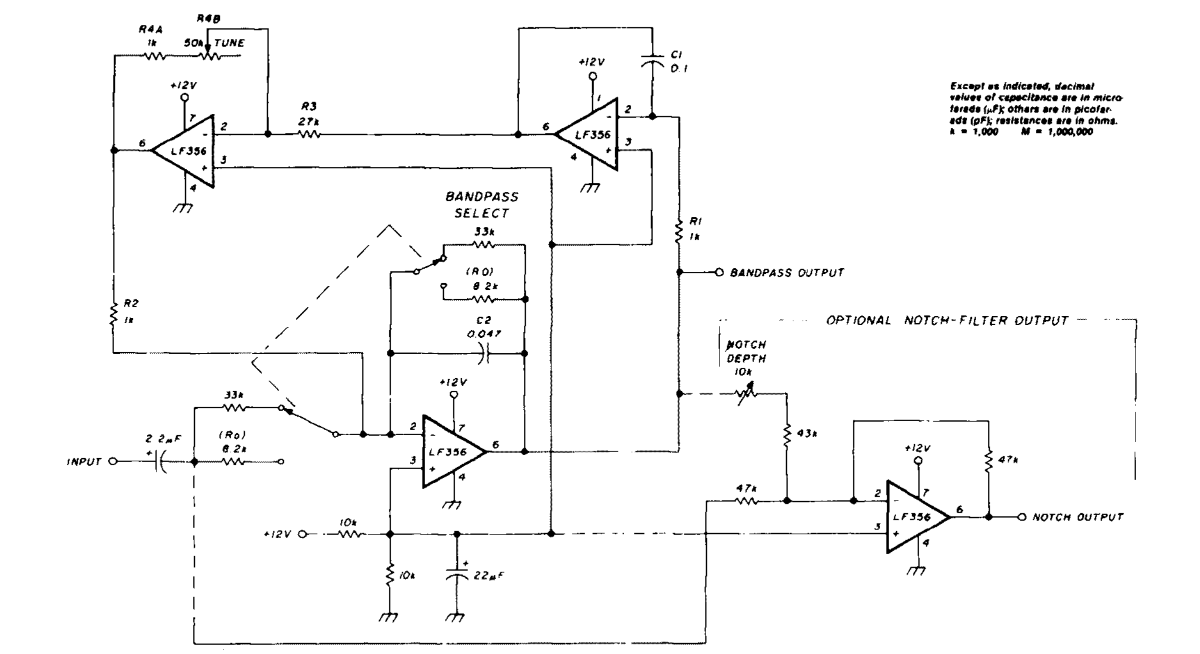


fig. 2. A variable-center-frequency and selectable-bandwidth filter is useful for CW reception.

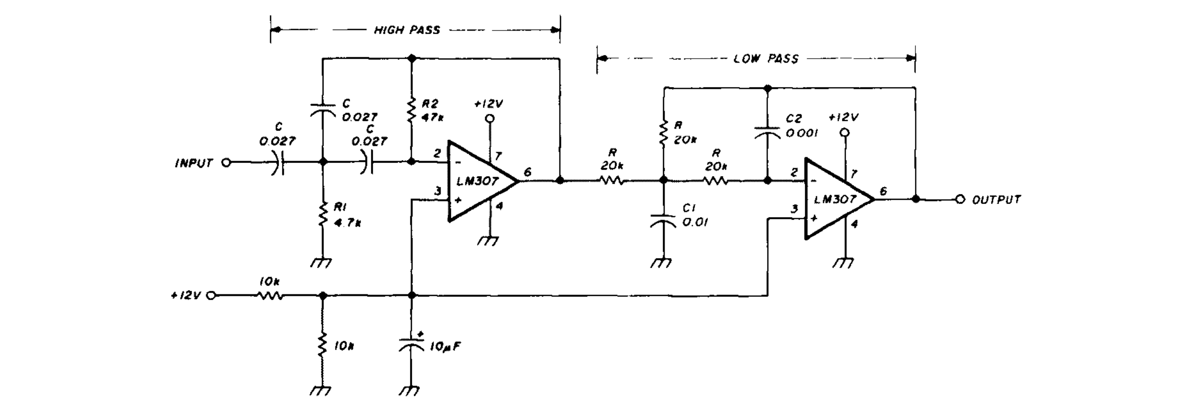


fig. 3. A combination 300-Hz highpass and 3-kHz lowpass filter for SSB use.

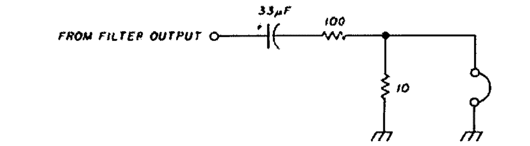


fig. 4. A circuit for coupling headphones to filter output.

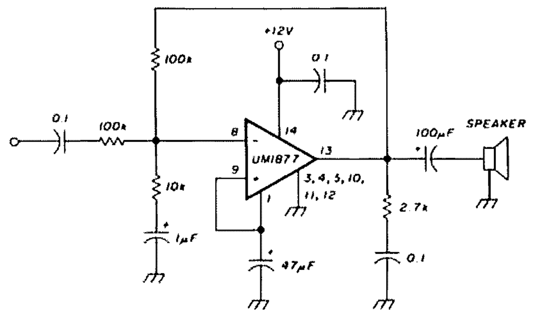


fig. 5. Unity-gain audio amplifier useful for driving a speaker from the filters.

passing the power supply near the chip, and all ground pins on the chip should be used.

summary

Now that you have the basic blocks, you can string them together to form a variety of filter combinations. A simple one-evening project can be made out of the bandpass filter, either with or without the notch output. I built the filter with simple perforated-board techniques and housed it in a small case. Add the SSB filter if you work phone and, of course, provide some means of switching the filters in and out. The audio amplifier is necessary only if headphones alone don't quite suit you. With unity-gain stages, the output should be the same level as the input, so if the audio is taken from a speaker or headphone jack the level can be easily adjusted with the receiver volume control to a usable level.

Please send me an SASE with any inquiries concerning this article. For more information on op-amps in general, see reference 3.

references

1. Aram Budak, "Passive and Active Network Analysis and Synthesis," Publisher Houghton-Mifflin, Boston, Massachusetts, 1974.
2. *Linear Databook*, National Semiconductor Corporation, Santa Clara, California, 1978.
3. Walter Jung, *IC Op-Amp Cookbook*, Howard W. Sams & Co., Indianapolis, Indiana, 1976.

ham radio

ALL ELECTRONICS CORP.

short circuits

power supply

In the article "Dual Voltage Power Supply" (*ham radio*, March, 1983) there is an error on the schematic on page 35. The two outputs of power supply A are tied together at the voltmeter connections. This should not be. Also, at the top of the same schematic, resistor R24 had been labeled R2H.

sideband transceiver

The following corrections should be made to the schematics and text of "15-meter Sideband Transceiver" (*ham radio,* March, 1983):

fig. 1: Change value of R26 from 100 to 10k ohms and value of R28 from 10k to 330 ohms.

fig. 6: Change component designations C66 to C60 and Q18 to Q24.

fig. 7: Add component values to R105 (100 ohms) and R106 (4700 ohms). R110 is a 2-watt resistor. Insert a resistor (R101, 330 ohms) in the collector lead of Q33.

fig. 8: Reroute emitter lead of Q25 to R88 and Q26 base junction. (It no longer goes directly to +10 volt bus.)

In the right-hand column on page 19, change component designations Q29 to Q20 and R66 to R67.

Be sure to check the artwork against the parts layout before beginning construction.

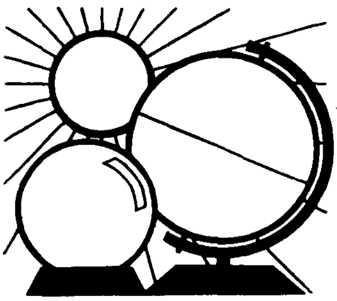
repeater antenna beam tilting

In K7NM's article, "Repeater Antenna Beam Tilting" (May, 1983), **eq. 2** should read as follows:

$$E_a = \frac{Sin\ n\left[(180^\circ s)\cos\theta + \frac{d}{2}\right]}{n\ Sin\left[(180^\circ s)\cos\theta + \frac{d}{2}\right]}$$

Eq. 4 should read this way:

$$A_h = \frac{0.0153P}{\sqrt{P}}$$



DX FORECASTER

Garth Stonehocker, KØRYW

last-minute forecast

The conditions this July will probably be considerably different from last year's. The summer months — normally a season of low maximum usable frequencies (MUFs) will bring even lower MUFs because of an advanced cycle smoothed sunspot number (SSN) as low as 60. Mid-latitude, zero-distance MUFs (f_oF_2-local noon) show a nearly linear variation with SSN, with 5.5 MHz, 8 MHz, and 11 MHz corresponding to SSNs of 10, 60, and 120 respectively.

July's forecast on the higher hf bands (10-30 meters) is for good long-skip conditions occurring the first and last weeks of the month and decreasing at other times. High and low latitude short-skip openings are expected to increase through sporadic E propagation during disturbed periods around the 5th, 10th, 21st, and 31st of the month. The lower bands (30-160 meters) should have the best nighttime DX during the in-between non-disturbed periods.

A full moon occurs on the 25th and perigee on the 11th of the month. The Aquarid meteor shower starts the 18th, peaks the 28th, and lasts until August 7th (all dates approximate). The radio-echo rate at maximum is about 34 per hour.

fading — QSA and QSB

Carefully observing daily DX signal levels will provide information on the state of the ionosphere and enable near future forecasting. Signal strength variations, fading, either decrease (attenuation) or increase (focusing), and possibly signal distortion will be heard. Fading is characterized by the duration of the interval between fades and the depth or decrease in amplitude of the signal during those periods. Most of the attenuation occurs as the signal travels through the D region (60-80 kilometer height) of the ionosphere. However, significant variations also occur at the area of reflection in the ionosphere, with signal levels modulated by geomagnetic field variations.

The following table lists four common types of fading conditions with the first two related to D region travel and the latter two occurring during layer reflection:

type of "fade"	cause	when/where	duration
SID	flare-ultraviolet and X-rays	daylight	1-2 hours
PCA	flare-proton particles	polar daylight	1-3 days
shortwave	solar wind-electrons (explained next month)	auroral zone (night)	2-5 nights
MUF failure	decreasing ionosphere (explained next month)	PM	½ hour

Solar radiation (ultraviolet and X-ray) produces D region absorption or attenuation, an attenuation that varies with the part of the sunspot cycle we're in, the time of year, and time of day. Signal level changes are slow and stable, except during solar flare induced sudden ionospheric disturbances (SID). These signal fades occur within 8 minutes on the sunlit propagation paths. The attenuation is a function of the cosine of the zenith angle to the sun. The typical time scale is a 10 to 20 minute decrease to maximum attenuation (lowest signal) and logarithmic return to the normal value within about one-half hour to two hours. The overall time (SID duration) is roughly related to flare size (importance or type) and radio flux (0.3 centimeter) burst shape and length.

Polar cap absorption (PCA) is also a D region slowly-varying attenuation effect produced inside of the auroral zone (polar cap) by protons arriving within an hour's time from certain solar flares. The attenuation is greater during daylight than at night. Therefore, the signal recovers somewhat each night then decreases during the day again, but shows improvement each day. The overall PCA attenuation duration is one to three days before normal propagation conditions are achieved again.

Both of these D region events occur mainly during the sunspot cycle peak and consequently should not bother us for a while. The shortwave fade and MUF failure are problems that can occur any time during the solar cycle and particularly during the solar cycle minimum. More about them next month.

band-by-band forecast

Ten and fifteen meters will have long-skip conditions in the afternoon during the peak times of the *27-day* solar maximum. Otherwise, look to sporadic E short-skip and multihop openings around *local* noon for DX on these bands. Transequatorial evening openings do not usually occur in the summertime.

JULY

WESTERN USA

GMT	PDT	N ↑ ASIA FAR EAST	NE ↗ EUROPE	E → S. AFRICA	SE ↘ S. AMERICA	S ↓ ANTARCTICA	SW ↙ NEW ZEALAND	W ← OCEANIA AUSTRALIA	NW ↖ JAPAN
0000	5:00	20	20	30	15	20	10	10	15
0100	6:00	20	20	30	15	20	10	10	15
0200	7:00	20	20	30	15	20	10	10	15
0300	8:00	20	20	30	15	20	10	10	20*
0400	9:00	20	20	40*	15	20	10	10	20
0500	10:00	20	20	30	20	20	10	10	20
0800	11:00	20	20	20	20	20	10	15	15
0700	12:00	20	20	20	20	20	15*	15	20
0800	1:00	—	20	20	20	20	15	15	20
0900	2:00	—	20	20	20	30	15	15	20
1000	3:00	—	—	30	30	30	15	20	20
1100	4:00	—	—	—	—	30	20	20	20
1200	5:00	—	—	—	—	—	20	20	20
1300	6:00	—	—	20	20	—	20	20	—
1400	7:00	—	20	20	15	—	20	—	—
1500	8:00	20	20	20	15	—	20	—	—
1600	9:00	20	20	20	15	—	20	20	—
1700	10:00	20	20	20	15	30	20	20	—
1800	11:00	20	20	20	15	20	20	20	20
1900	12:00	20	20	20	15	20	15	20	20
2000	1:00	20	20	20	15	20	15	15	20
2100	2:00	20	20	20	15*	20*	15*	15	20
2200	3:00	20	20	20	15*	15	15*	15*	20
2300	4:00	20	20	20	15	15	10	15*	20

MID USA

GMT	MDT	N ↑ ASIA FAR EAST	NE ↗ EUROPE	E → S. AFRICA	SE ↘ S. AMERICA	S ↓ ANTARCTICA	SW ↙ NEW ZEALAND	W ← OCEANIA AUSTRALIA	NW ↖ JAPAN	CDT
0000	6:00	20	20	30	15	20	15*	15	15	7:00
0100	7:00	20*	20	30	15	20	15*	15*	20*	8:00
0200	8:00	15	20	30	15	30	15*	15*	20	9:00
0300	9:00	20	20	30	15	30	15	15	20	10:00
0400	10:00	20	20	40*	15	30	15	15	20	11:00
0500	11:00	20	20	30	20	30	15	15	20	12:00
0800	12:00	20	20	20	20	30	15	15	20	1:00
0700	1:00	—	20	20	20	30	15	20	20	2:00
0800	2:00	—	20	20	20	30	20	20	20	3:00
0900	3:00	—	20	20	20	30	20	20	20	4:00
1000	4:00	—	20	20	20	30	20	20	20	5:00
1100	5:00	—	—	20	20	—	20	20	—	6:00
1200	6:00	—	—	—	20	—	—	—	—	7:00
1300	7:00	—	—	—	—	—	—	—	—	8:00
1400	8:00	20	20	20	15	—	20	—	—	9:00
1500	9:00	20	20	20	15	—	20	20	—	10:00
1600	10:00	20	20	20	15	—	20	20	—	11:00
1700	11:00	20	20	20	15	20	20	20	—	12:00
1800	12:00	20	20	20	15	20	20	20	—	1:00
1900	1:00	20	20	20	10	20	15	20	20	2:00
2000	2:00	—	20	20	10	20	15	15	20	3:00
2100	3:00	—	20	20	10	20	15	15	20	4:00
2200	4:00	—	20	20	15	20	15	15	20	5:00
2300	5:00	—	20	20	15	20	15	15	20	6:00

EASTERN USA

GMT	EDT	N ↑ ASIA FAR EAST	NE ↗ EUROPE	E → S. AFRICA	SE ↘ CARIBBEAN S. AMERICA	S ↓ ANTARCTICA	SW ↙ NEW ZEALAND	W ← OCEANIA AUSTRALIA	NW ↖ JAPAN
0000	8:00	20*	20	30	15	30	15	15	20
0100	9:00	20	20	30	15	30	15	15	20
0200	10:00	20	20	30	15	30	15	15	20*
0300	11:00	20	20	30	15	30	15	15	20
0400	12:00	20	20	40*	15	30	15	15	20
0500	1:00	20	20	30	15	30	15	20	20
0800	2:00	—	20	20	20	30	20	20	20
0700	3:00	—	20	20	20	30	20	20	20
0800	4:00	—	20	20	20	30	20	20	20
0900	5:00	—	20	20	20	30	20	20	20
1000	6:00	—	20	20	20	—	20	20	—
1100	7:00	—	—	20	20	—	20	—	—
1200	8:00	—	—	—	20	—	—	—	—
1300	9:00	20	—	15	15	—	—	—	—
1400	10:00	20	20	15	15	—	20	—	—
1500	11:00	20	20	15	15	—	20	20	—
1600	12:00	20	15	15	15	30	20	20	—
1700	1:00	20	15	15	15*	20	20	20	—
1800	2:00	20	15	15	10	20	20	20	—
1900	3:00	—	15	15	10	20	20	20	20
2000	4:00	—	15	15	10	20	15	15	20
2100	5:00	—	20*	20	10	20	15	15	20
2200	6:00	20	15	20	15*	20	15	15	20
2300	7:00	20	15	20	15*	20	15	15	20

*Look at next **higher** band for possible openings.

Twenty and thirty meters will be open all day and much of the night. If twenty does not stay open through the night, thirty probably will. Sporadic E short-skip is also often effective on these bands throughout the day. Propagation paths to most areas of the world are viable in a sequence that follows the sun's journey across the sky: east in the morning, south during mid-day, and west during the evening.

Thirty and forty meters will be the main nighttime DX bands this time of year, though long-skip distances will be shorter. Sporadic E openings are possible during more of the day into pre-sunrise and after sunset. With thunderstorm-induced static levels high in the evening, look to pre-dawn periods for best results.

Eighty and one-sixty meters are difficult DX bands this time of year. Short nights and high noise levels hamper DX operation with eighty having slightly lower noise levels. Most useful openings may occur during the pre-dawn hours. Sporadic E propagation signal strengths may exceed the static level near sunrise and sunset.

ham radio

technical forum

Welcome to the ham radio *Technical Forum. The purpose of this feature is to help you, the reader, find answers to your questions, and to give you a chance to answer the questions of your fellow Radio Amateurs. Do you have a question? Send it in!*

Each month our editors will select the best answer received to a question previously posed in Technical Forum. We'll send the writer a book from our Bookstore as a way of saying thanks.

measuring inductances

In February, 1983, Technical Forum published a request from K9EBA for information on the measurement of low values of inductance.

Several years ago the San Bernardino Microwave Society addressed this problem and came up with a simple circuit for measuring small values of inductance. It was published as a NASA Tech Brief. This circuit used the parts on hand at the time. The circuit works well and has been duplicated by several experimenters. It measures inductances between 30 nH and 30 μH. This is not the only way the circuit can be implemented, nor even the best way, but it is one method that works.

The only trick in building the circuit is to minimize the stray shunt capacity across the unknown inductance. I used a 1-inch hole, with a 4-40 (M3) screw in the center and a thin sheet of plastic to support it. Fiber shoulder washers for the unknown terminal have too much stray capacity, but other than this, the circuit is straightforward and should pose no problems. — **Richard B. Kolbly, K6HIJ.**

Ed. note: An SASE to *ham radio* will bring the interested reader a copy of the NASA Tech Brief and associated technical support package describing the direct-reading inductance meter.

impedance matching

I wound an rf impedance matching transformer on an iron powder toroid core (T225-2 mix) for a 50-ohm to 300-ohm transformation. I used a turns ratio of just under 2.5 to 1; that is, I wound seventy-three turns of No. 20 Formvar enamel wire next to the toroid core (300-ohm winding) and thirty turns of No. 16 Teflon-covered wire on top of it (50-ohm winding). There is more than one inch of empty core space between the ends of the high-impedance winding. The thirty turns of the low-impedance winding are centered over the middle of the seventy-three-turn winding. It is wound in the same *direction and* covers about half of the circumference of the toroid.

I tried to feed a few watts of rf power into a 300-ohm carbon resistor attached to the 300-ohm winding as a test on 29 MHz. It failed completely. It would not load up and had an SWR of over 10:1. I then checked the impedance of the low-impedance winding with an rf noise bridge (with the 300-ohm resistor still connected to the seventy-three-turn winding). I found that the impedance was indeed between 50 and 60 ohms resistive, but it had a very high capacitive reactive component of 60 to 70 pF.

Does anyone have any explanation of this result? — **Joseph Neiman, WB2NTQ.**

static mystery

Over the past thirty-seven years of shortwave listening I have observed a steady increase of that hammering and hissing noise called "rain static." I do not remember a single incident of this phenomenon while operating in Switzerland from 1946 to 1948.

The first time I encountered it was in late 1948 in the vicinity of Cleveland, Ohio. At the time I guessed that the Cleveland weather conditions might be somehow different from Swiss weather conditions.

Through 1949 and 1950 I got used to rain static in New Jersey. When I returned to Switzerland I found things quiet again no matter how heavy the rain. But by about 1955 I began to notice subtle signs of Swiss rain static which appeared, through the years, more frequently and more intensively.

At present about forty percent of all medium-strength rainfalls here cause rain static, and the amount seems to be increasing.

It is known that split water droplets can become charged, probably by a kind of tribo-electric effect. If such droplets hit antenna elements, charge compensation by the antenna could account for the observed receiver noise. So the question remains, why was the effect not observed in Switzerland before 1955, but already encountered in Ohio by 1948 and in New Jersey shortly thereafter?

Could there be some connection with air pollution caused by industry and automobile traffic, thus enhancing charge separation of water droplets?

Not knowing enough about electrostatics and electrochemistry, let me present this problem to you and your readers in the hope that someone might provide a physical model or references to published work.

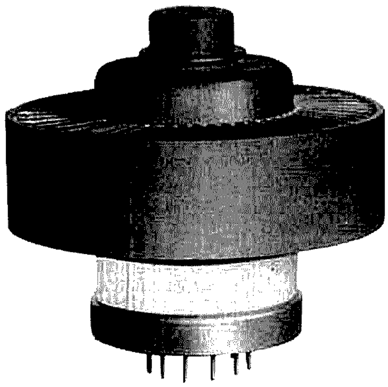
Are there any effective countermeasures which could eliminate this kind of interference? — **Bruno Binggeli, HB9FU.**

NEW products

power triode

The EIMAC Division of Varian has announced the availability of a new ceramic/metal power triode intended for use as a cathode-driven amplifier for hf and vhf service. This compact tube (3CX800A7) is intended for high power linear amplifier service. A single tube will produce a full 2 kW PEP or 1 kW CW input power.

The rugged 3CX800A7 is rated for 800 watts plate dissipation and will deliver full power output with less than 40 watts peak drive power. Power gain is better than 15 dB. The air-cooled anode requires less than 20 cfm with a backpressure rating of 0.35 cfm for full dissipation *at sea level.*



Height of the 3CX800A7 above the socket plane is only 2-1/4 inches (5.7 cm), making the tube well suited for compact linear amplifier design and compatible with modern, low-profile styling.

For further details, contact Varian, EIMAC Division, 301 Industrial Way, San Carlos, California 94070. RS#301

Editor's Note: *Both Henry Radio and Ehrhorn Technological Operations (ETO) have designed new amplifiers around this new tube. Contact them for details*

polarization control

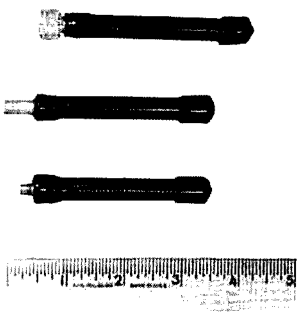
TEM Microwave Corporation is pleased to announce its model SC-10 polarization control interface. The SC-10 is designed to interface with satellite TVRO receivers that have odd/even channel logic output signals, such as the R.L. Drake ESR-24, or SPDT contacts, such as the Automation Techniques GLR-500 series. The SC-10 produces the correct power and drive signals to control the popular servo motor type feed systems, such as the Chaparall Polarotor 1™ or the Boman EFH-75. Other features of the SC-10 include independent front panel horizontal and vertical fine adjustment control and LED indicators that show which control is enabled, a mode switch for choice of either Satcom or Westar-type polarization, and a built-in regulated power supply.

The size is 4 × 5 × 2 inches (10.16 × 12.7 × 5.08 cm). Power is UL listed plug-in wall transformer. For more information, contact TEM Microwave Corporation, 22518 - 97th Avenue North, Corcoran, Minnesota 55374. RS#302

1/4-wave replacement antennas

Centurion International, Inc., has introduced a new line of 1/4-wave, flexible, miniaturized replacement antennas for VHF frequencies. The new "style-S" antennas measure approximately 3 inches in length by 3/8-inch in diameter. These antennas are smaller in diameter than other 1/4-wave miniaturized antennas and are more flexible. Their reduced size makes *them a good choice* for use with smaller portable two-way radios and speaker microphones.

Designated the "Skinny Mini," the antennas are encapsulated in high-gloss PVC and remain flexible from −55C to 100C. Style-S antennas, like style-M, are available with any of more than twenty different base connector configurations, to fit virtually any radio made.

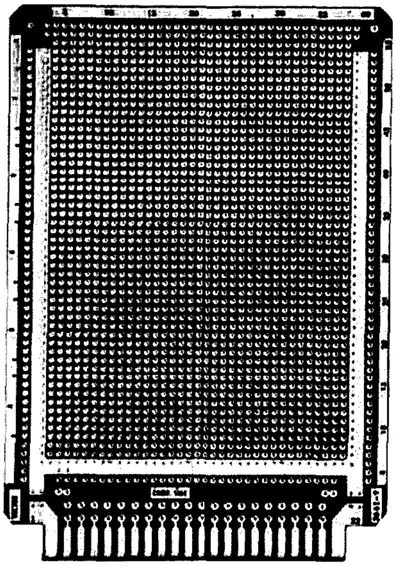


For more information, contact Centurion International, Box 82846, Lincoln, Nebraska 68501. RS#303

plug-in circuit boards

Three new plug-in circuit boards from Vector Electronic Company incorporate individual solder pads and drilled, plated-through holes. The design allows complete freedom in component location and spacing while providing quick and easy solder mounting of components with solderable or wrap-post leads. The boards have 2064 holes in the component area, allowing placement of up to fifty fourteen-pin DIPS or forty sixteen-pin DIPS. One card, the Model 4610-3, is form and fit compatible with STD system cards with 28/56 card-edge contacts. The Model 3662-9 and Model 3619-6 have 22/44 and 36/72 card-edge contacts to mate with the most frequently used connectors.

All boards are 4.5 inches wide by 6.5 inches long by 0.062-inch thick (11.43 × 16.51 × 0.16 cm) and have 0.042-inch (0.107-cm) diameter plated-through holes on 0.1-inch (0.25-cm) centers.



Fabricated of FR-4 (G10) epoxy glass laminate, the pads are 2-ounce copper cladding with bright tin plating for easy soldering. Card-edge connectors are nickel plated and gold flashed to ensure long life and low resistance. Zoned-wiring locations, etched into the cladding, permit easy component identification.

In single quantities, the 22/44-contact Model 3662-9 is priced at \$26.80 each; the 28/56 contact Model 4610-3 is \$26.50 each; and, the 36/72 contact Model 3619-6 is \$26.80 each. For more information, contact Vector Electronic Co., Inc., 12460 Gladstone Avenue, Sylmar, California 91342. RS#304

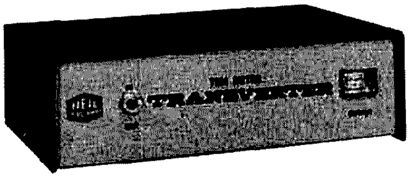
Say You Saw It In HAM RADIO

NEW products

ten-meter fm transverter

A unique 2-meter to 10-meter linear translator recently introduced by Heil, Ltd., allows a 2-meter radio to receive and transmit on the ten-meter band from 28.00 to 29.70.

The Model 210 is primarily designed for use in the 29.30 to 29.70 fm band using a one-watt "handie talkie" or mobile transceiver for excitation, but is also usable on SSB, CW, a-m, and RTTY by exciting with an all-mode two-meter rig. The Model 210 has three SO-239 connectors on the rear panel, a two-meter one-watt input, a two-meter antenna, and a ten-meter antenna. With the front panel function switch in the "out" position, the two-meter antenna is connected to the two-meter transceiver or "handie talkie." Switching to the "in" position will cause the transverter to operate and produce a signal in the ten-meter band. The receiver sensitivity is 0.3 μV for 10 dB quieting. The output power is approximately 4 watts out at 29.60.



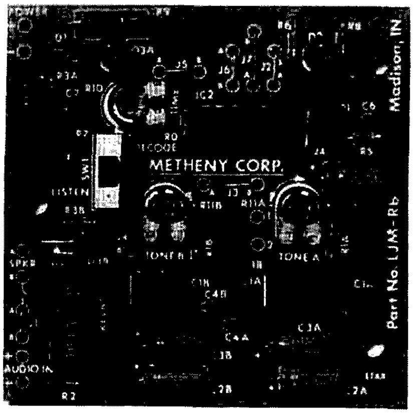
The price (subject to change) is $100.00. For further details, write Heil Sound System, Heil Industrial Blvd., Marissa, Illinois 62257. RS#305

emergency tone decoder

The Storm Alert LJM2RK time-dual tone emergency decoder kit from Metheny converts receivers into special-purpose receivers or controls. When a user-selected time-tone combination is received, the output provides a relay control for activating speakers or other devices.

Special features include single or dual tones adjustable over the touch tone range; adjustable time delay; relay output; manual or auto reset; single tone ON latching with different single tone reset OFF; and interfacing of multiple boards for multi-digit sequential activation and reset.

Kit LJM2RK includes a printed circuit board with components, relay, and a silk screened component identification and solder mask for ease of assembly. An optional enclosure kit, LJM2RC, includes a custom-molded case, speaker, audio input cable, and hardware for



the decoder kit. Kit LJM2RK costs $15.00; the enclosure kit, LJM2RC, is priced at $5.00.

For complete details and information about specific applications, contact The Metheny Corporation, 204 Sunrise Drive, Madison, Indiana 47250. RS#306

multimode transceiver

The FT-726R — the world's first Amateur HF/VHF/UHF transceiver capable of full duplex operation for satellite work — is now available from Yaesu Electronics Corporation.

The basic unit comes equipped for 2-meter operation on SSB, CW, and fm. Optional units may be plugged in, enabling operation on 10 or 6 meters, 430 to 440 or 440 to 450 MHz on 70 cm. The optional SU-726 satellite unit allows crossband full duplex operation, for simultaneous uplink transmit and downlink receive operation on Amateur satellites.

Controlled by an eight-bit microprocessor, the FT-726R features a dual VFO and memory frequency management system, with independent frequency/mode storage on each VFO or memory; mode-inverting satellite transponders are therefore covered with ease. The transmit and receive frequencies may be varied during satellite work to allow easy zero-beat capability while following Doppler shift.

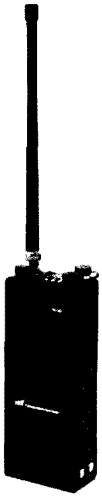
Equipped with many features found only on

hf transceivers, the FT-726R includes an SSB speech processor, i-f shift, variable i-f bandwidth tuning, i-f noise blanker, RIT, multimode squelch, and a receiver audio tone control. A CW filter, DTMF encoding microphone (YM-48), desk microphone (MD-1B8), external speaker (SP-102), and CTCSS units are all available as options.

For further information, contact Yaesu Electronics, P.O. Box 49, Paramount, California 90723. RS#307

handheld airband transceiver

The TR-720 is a solid-state, fully synthesized, portable airband transceiver covering the 720 COM channels between 118 and 136 MHz and 200 NAV channels from 108 to 118 MHz. It measures only 6.6 × 2.6 × 1.5 inches and weighs just 19 ounces. It employs microprocessor technology, has a twist-off battery pack, comes with a complete set of accessories, is FCC type accepted, and carries a full one year warranty. It is available for $795.00 from local Avionics dealers, or directly from the manufacturer.



For information, contact Communications Specialists, Inc., 426 W. Taft Ave., Orange, California 91665. RS#308

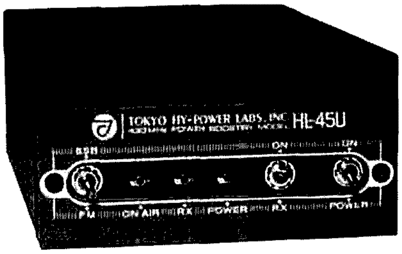
power bars

A new line of Hammond power bars features an attractive, contemporary, brushed-aluminum case with matte black receptacle housing. Reduced in size, (11, 14 and 17 inches in length), standard models are available in four, six, or eight-receptacle sizes with either 6 or 15 foot cords, and with or without lighted, rocker type on/off switches. Also available are 4 and 6 foot long power bars, each with eight receptacles. Appropriate for work station mounting, all power bars are CSA approved and fitted with 120 Vac, 15A circuit breaker.

For more information, contact Hammond Manufacturing Company, Inc., 1690 Walden Avenue, Buffalo, New York 14225. RS#309

uhf linear amplifier

The newest uhf linear amplifier from Tokyo Hy Power Labs is designed for use with 10-watt output 430-450 MHz crystal-controlled or synthesized rigs. Input for the HL-45U is 2 to 15 watts with output of 10 to 45 watts. It operates from a 13.8 volt dc source and draws 7 amps at 45 watts output. It is all-mode (SSB, CW, and fm), has a built-in 12 dB low noise receiver pre-amplifier, and employs carrier operated switching (COX).

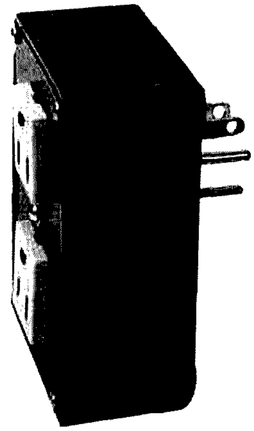


The HL-45U measures 4.9 × 2.7 × 6.7 inches (124 × 68 × 170 mm) and weighs 2.76 pounds (1.25 kg). The suggested retail price of the HL-45U is $199.95.

For further information contact THL Sales Department, Encomm, Inc., 2000 Ave. G, Suite 800, Plano, Texas 75074. RS#310

wall socket RFI control

Electronic Specialists' new Direct Plug Super Filter and Suppressor features a dual-pi filter to control electrical interference. A 6500-ampere spike/surge suppressor protects equipment from damage caused by lightning or heavy machinery spikes.



For further information, contact Electronic Specialists, Inc., 171 S. Main Street, Natick, Massachusetts 01760. RS#311

ham radio magazine

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- testing baluns
- digital audio filter for CW and RTTY
- hybrid couplers
- antenna carriage and track pole mount

focus on communications technology

RF SYNTHESIZERS

ham radio

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AUGUST 1983

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contents

REFLECTIONS

SPECIAL REPORT: W5LFL and SPACELAB STS-9

A three-hour teleconference call on June 9 brought together representatives of NASA, AMSAT, the broadcasting industry, and several key figures in Amateur Radio publishing. The subject: the 9-day STS Spacelab Mission of Dr. Owen Garriott, W5LFL. While this page is normally reserved for the editorial — and was scheduled this month to include the 1983 ham radio *Reader Survey — we gladly postpone both in order to bring you details of the project.*

For the first time in the history of the space program, individual citizens will be able to talk directly with an astronaut *in space* when NASA mission specialist Dr. Owen Garriott, W5LFL, attempts to contact as many Amateurs worldwide as possible during the last five days of the STS-9 spacelab mission, scheduled for launch on September 30.

W5LFL will be able to operate only during his leisure time periods, expected to be one hour per day, for a mission total of five hours. He will transmit on the even minutes for one minute and listen on the next odd minute while logging and tape recording what he hears. He will then acknowledge call signs heard on the next even minute. It is expected that he will QSO approximately 500 hams worldwide and be heard by 300,000. The ARRL will act as QSL manager and provide acknowledgements for all verified QSOs and SWL reports.

The equipment aboard the spacecraft will be a multichannel black box transceiver capable of 5 watts output from its battery power. The station, located on the aft flight deck of the space shuttle orbiter Columbia, will use a printed circuit loop antenna mounted on the upper crew compartment window.

Operation will take place on uplink frequencies (earth to space) of 144.910-145.470 (possibly no higher than 145.190) in 20 kHz steps while the downlink frequencies (space to earth) will be 145.510-145.770 (2 or 3 specific frequencies will be chosen), also in 20 kHz steps. Mode of operation will be 2-meter fm (split).

The 57-degree inclination, 90 minute, 155 mile altitude orbit will enable most of the earth's land mass to be (line-of-sight) visible from the spacecraft during a typical day, though any specific location will have an 8-minute maximum pass.

AMSAT recommends that ground station equipment consist of no more than a 10-15 watt 2-meter transceiver feeding a high lobe (80-90 degrees would be ideal) turnstile antenna mounted above any obstructions. They do not recommend using more elaborate antennas (such as a cross-Yagi).

A combined effort on the part of many individuals and organizations,* this project is the result of a joint proposal by the ARRL and AMSAT to NASA. Earlier proposals to place an Amateur Radio transceiver aboard an orbiting U.S. spacecraft date back to the early '70s, when a project called "SKYLARC" — Skylab Amateur Radio Communications — was planned but scrubbed because it came too late in the development of the program. The present proposal was recently accepted by NASA, with the only restrictions related to non-interference with higher priority mission objectives, systems and, of course, safety.

Starting in mid-July the ARRL is expected to provide a 900 telephone number for the latest orbital information. Other sources of information are NASA itself, which will provide a timetable (flightline) of the operation, and Westlink, whose report can be heard by calling (213) 465-5550.

The ARRL is also planning a videotape presentation featuring Dr. Garriott. Hosted by NBC's Roy Neal, it will document the role of Amateur Radio and Amateur Radio operators throughout the STS-9 mission.

This is truly a unique occasion for Radio Amateurs to show the world how they can contribute to technology and public awareness. Each Amateur will provide quite some service if he is able to acquire the signal, tape record it and contact the local media with details of the QSL and perhaps even a recording of any voice contact. Reporters may want to play the tape over their broadcast stations or transcribe it for newspaper use.

So clean up the shack, tape record everything you hear, and *good luck*!

Rich Rosen, K2RR
Editor-in-Chief

***credit where it's due**

A short list of just some of the people instrumental in bringing about this effort is definitely in order: General James Abramson, Assistant Administrator of NASA, who gave final approval for Amateur operation on the mission; Bernie Glassmeyer, W9KDR, ARRL Space Program Manager; Peter O'Dell, KB1N, ARRL Public Information Coordinator; Steve Mendelsohn, WA2DHF, CBS technician and Vice-Director of the Hudson Division of the ARRL; Rich Moseson, N2BFG, Associate Producer, CBS News; Bill Tynan, W3XO, VHF Contributing Editor, QST; Roy Neal, K6DUE, NBC Science Editor; Vern Riportella, WA2LQQ, AMSAT President-Elect; and Bill Pasternak, WA6ITF, Editor, Westlink Report.

presstop de W9JUV

FCC'S "NO MAILBACK" NOVICE EXAM PROPOSAL WAS APPROVED June 29 by the Commissioners and goes into effect by the end of August. Under the new procedure, proposed in PR-Docket 82-727, a General Class or higher Amateur will make up a Novice exam from a bank of 200 FCC approved questions, then both administer and grade it himself. He'll then simply advise FCC's Gettysburg office if the applicant passes, and they'll issue a license.

The New Procedure Should Cut Licensing For A Novice to a fraction of its current turnaround time, and save the Commission a good deal of money as well. The list of 200 questions and procedural detail should be ready for distribution by mid-July.

"Business Communications" Was Another Issue Addressed at that same meeting, resulting in the adoption of an "Interpretive Order." A new paragraph has been added to Rules Part E (Prohibited Practice) which states simply that business communications are prohibited and then defines business communications as "...any transmission...to facilitate the regular business or commercial affairs of any party." which is the definition now found in (Amateur section) Part 97.114(c). Its application was also specifically extended to overseas third-party traffic.

FCC's Proposal To Change Power Limits And Measurements from dc input to rf output may also make it to the Commissioners' agenda before they adjourn for August recess.

The Ancient Issue Of Amateur "Broadcasting" may also turn up on the Commission agenda in the near future. Most recently at issue was the rebroadcast of Space Shuttle communications, but the question of what constitutes "broadcasting" in the Amateur Service and when an Amateur "broadcast" is justified has never been formally addressed by the FCC.

Informal Comments And Suggestions On The Broadcasting question could prove helpful at this time, when the issue is still in the discussion stage. Contact Personal Radio Branch Chief John Johnston or Private Radio Bureau Chief James McKinney.

CONTACTS WITH W5LFL/SPACE DURING SKYLAB'S OCTOBER VOYAGE will necessarily be quite limited, with no more than 500 lucky Amateurs likely to hear their calls to space acknowledged by Astronaut Owen Garriott as he passes overhead. An operating procedure has already been established (see this month's editorial for this and other technical information).

Final Details Will Become Available as the late September launch date approaches. Check ARRL bulletins, AMSAT nets, and Westlink's 213-465-5550 line after mid-September.

OSCAR 10 IS UP AND LOOKING GOOD, following a textbook Ariane launch at 1159Z June 16. 17 minutes into launch OSCAR 10 separated, entering a highly elliptical temporary orbit. Due to a less than ideal attitude the first thrust motor burn was delayed until the end of June; the last burn (to final orbit) should be mid- to late-July, after which the Mode B transponder will be turned on. Model L (1268 to 436 MHz) won't be turned on until later.

OSCAR 10's 145.810 MHz Beacon Has Been Putting Out an excellent signal, with a five minute CW "status report" every hour and half hour and telemetry data in between. Due to the 200 by 39,000 kilometer elliptical orbit, the beacon has been audible in the midwest for more than 8 hours at a time with slowly changing azimuth and elevation bearings.

THE FUTURE OF THE VOLUNTEER AMATEUR EXAM PROGRAM has been clouded by an apparent ARRL policy change incorporated in its Reply Comments on the exam program NPRM, PR Docket 83-27. Their new position, which is reported to have taken the Commission by surprise, is that the League wouldn't participate in the volunteer program unless it was permitted to recover costs of operating the program by charging applicants an examination fee.

Sen. Barry Goldwater, K7UGA, Whose Communications Act Rewrite provided the means for setting up the volunteer program, has just told WA6ITF of Westlink that he is completely opposed to any such fee. Goldwater does, however, support the League against "No-Code."

AMSAT WILL OPPOSE OPENING 29-29.5 MHZ TO REPEATERS, as proposed in FCC's PR Docket 83-485. Comments were due July 25, and Reply Comments will be due August 24.

ALL AMATEUR LOG KEEPING REQUIREMENTS HAVE BEEN DROPPED by the FCC, including those concerning logging of repeater autopatches and even overseas third-party traffic. Though an Amateur may still keep a log for his own use, it's no longer a Part 97 requirement.

However, Some Related Information Previously Required under Part 97.103 must still be kept but has been moved to other parts of the rules. Examples include much of the information pertaining to remotely controlled stations and their security.

In A "Streamlining" Move, The FCC Has Also Adopted a number of other rules changes. A station operating ATV or RTTY, for example, now need identify only in the mode in which he's operating. An interim permit may now be renewed, and the rules now spell out the right of the Commission to inspect an Amateur station.

ERIC SHALKHAUSER, W9CI, PASSED AWAY June 9. He was one of the few remaining active Amateurs from before WWI and among his many accomplishments were his designs of the RME 69 and 70 plus many other outstanding receivers and converters for Amateurs.

comments

no code

Dear HR:

Isn't all this wailing against a "no-code" license a little illogical?

Code became a licensing requirement because all radio communication then was by code. Today, almost none is. Only a few Amateurs and a handful of outmoded commercial and maritime services use CW and code.

In today's hyper-electronic age, requiring code proficiency in order to operate a radio-telephone (which is all most hams want to do) is like making you learn to ride a motorcycle well in order to get a license to drive an automobile. Silly, eh?

Some people like motorcycles. They're fun, I'm sure. And they get you where you want to go, like automobiles do. But there the resemblance ends. The number of motorcycles in use probably compares with the number of automobiles in use in about the same ratio as the number of active CW stations compares with the number of active SSB stations. CW is fun, and I was a "CW Forever" ham for fifty-four years until I was given a mike. And CW provides communications, like SSB. But again — there the resemblance ends.

Once, you had to know code to use radios. Today you don't. The great majority of newly licensed hams forget the code as soon as they get on the air. Many never even own a telegraph key.

It would seem that those hams most outspoken against the "no-code" license simply feel that everybody should suffer like they did *because* they did. Human, perhaps, but illogical.

Forcing people to learn code has not filled our CW bands. Count the CW signals any time of the day between 14000 and 14200, and the SSB carriers between 14200 and 14350 and see how they compare. Sad, isn't it? Do we want to lose all that unused space to some other (and more pressing) service — and we will, you can bet on it — or stop pushing an anachronism and let the 'phones use it for expansion, as is now happening?

To paraphrase an old saying, "You can force a ham to learn the code, but you can't make him use it." Let's get smart.

Bill Lippman, W6SN
Pacific Palisades, California

converting TV

Dear HR:

The article by Carl Gregory, K8CG, (*ham radio*, April, 1983) on converting TV sets to video monitors for computer use was well done and informative, but he missed an important (and very simple) method of improving performance.

A couple of years ago I found myself in need of a video monitor for computer and high resolution VCR use and I too decided to convert a TV set. Don Lancaster covers the subject beautifully in Chapter 8 in his *TV Typewriter Cookbook*. Following his instructions, I converted a Motorola 20TS chassis. It worked well, but showed the same defects as K8CG's "simplest approach" solution — mainly, blurred pixels in computer use.

The situation was vastly improved by simply including another Lancaster suggestion: shorting out the 4.5 MHz sound trap in the output of the final video amplifier. As long as the sound trap is in the circuit, the set's video response has a big hole in it. With the short rise times in the computer generated signal, such a gap in response can be fatal.

I will not deny that K8CG's approach using an external video amplifier tied directly to the CRT cathode is superior (especially in terms of contrast and the best possible video bandwidth), but for those of us who are essentially too lazy to go that route, the sound trap refinement will make the easier route adequate for most purposes. Both the Betamax and TRS-80 Model I seem to be happy with the simpler approach, and so am I.

Tom Adams, K9TA
Marinette, Wisconsin

CATVI

Dear HR:

After seeing all the articles written in various magazines last year on CATVI, I did not think it could happen to me. I felt an underground cable system should be virtually problem-free.

I am now involved with a complaint to the FCC and writing this letter to warn other hams one more time. You need to keep a constant surveillance on your cable system for leaks. One of the easier frequencies to watch is 145.250 and 144.0 if you're on an HRC system. To be sure of the frequencies used, contact your local cable office and ask the engineer in charge. A new cable installation can easily cause TVI to your entire neighborhood if it is not installed correctly. Also keep in mind that your cable system does not *have to use a ham* band frequency before you can cause interference to a cable system.

Ron Hooper, WB4NMA
Gainesville, Georgia

Demystifying
PLL synthesizer design

rf synthesizers for high-frequency communications: part 1

PLL frequency synthesis has become the dominant frequency generation technique in modern communications equipment as a result of the increased availability of PLL oriented ICs. This, in turn, has enabled the design of compact-low-power, inexpensive synthesizers. While this technology has been accessible for some time, it's necessary to understand how PLL synthesizers work before trying to design or build one.

the phase-locked loop

There are three basic circuit blocks in a PLL: the phase detector, the loop filter, and the voltage-controlled oscillator (VCO), as shown in **fig. 1**.

In a PLL two signals at the phase detector input are in phase and therefore matched in frequency. Should they not be in phase, the phase detector generates an error signal which is filtered and passed to the VCO to correct the phase difference. At all times the loop tries to maintain zero phase difference between the phase detector input signals. A synthesizer is formed through the addition of a frequency divider to the basic PLL, (**fig. 2**). A zero phase error at the phase detector input occurs when the VCO is operating at a frequency

$$f_{VCO} = Nf_{REF}$$

Since $\div N$ is a programmable divider, the VCO can oscillate at any frequency within its range as long as that frequency is a multiple of f_{REF}, simply by changing N.

Fig. 3 illustrates a synthesizer designed to tune from 5.000 to 5.500 MHz in 1 kHz steps, using a 1 kHz reference oscillator and a programmable divider with a range of 5000 to 5500.

phase detector

The phase detector can be built around either analog or digital circuitry. However, since it is easier to design the PLL around digital phase detectors than around analog phase detectors, only the digital phase detectors will be discussed.

Digital phase detectors use gates and flip-flops to detect phase and frequency differences. Their output are pulses whose average value is a dc voltage that is dependent on phase difference. A plot of voltage output versus phase difference input of a phase detector is shown in **fig. 4**.

From **fig. 4**, phase detector gain, K_ϕ, is defined as:

$$K_\phi = \frac{\Delta V_{OUT}}{\Delta\theta_{IN}}$$

where ΔV_{OUT} is the averaged phase detector output voltage change and $\Delta\theta_{IN}$ is the input phase difference in radians.

By Craig Corsetto, WA6OAA, 4312 Marlowe Drive, San Jose, California 95124

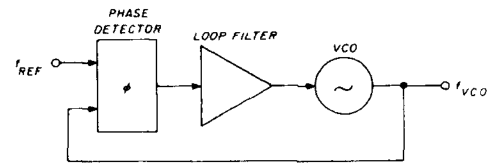

fig. 1. Basic PLL circuit blocks.

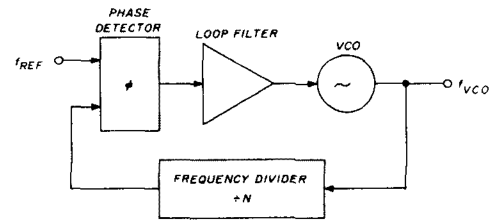

fig. 2. Circuit blocks for simple PLL synthesizer.

In **fig. 4**, K_ϕ is:

$$K_\phi = \frac{5V - 0V}{2\pi - (-2\pi)} = \frac{5V}{4\pi}$$

$$\approx 0.397 \text{ volts/radian}$$

K_ϕ is usually given in the manufacturer's specifications or in the applications notes for the device.

VCO

The VCO is an oscillator whose frequency is controlled by an externally applied voltage, V_T. For good phase noise performance the VCO usually employs an LC or crystal resonator because of the high **Q** of these components.

The VCO has a gain term, K_{VCO}, which is the amount of frequency change, Δf_{VCO}, caused by a change in voltage V_T, on the VCO tune line.

$$K_{VCO} = 2\pi \frac{\Delta f_{VCO}}{\Delta V_T}$$

Since varactor diodes have nonlinear capacitance versus voltage characteristics, K_{VCO} is not constant over the VCO frequency range. Therefore, two values of K_{VCO} should be specified at the two ends of the VCO range.

Fig. 5 illustrates how f_{VCO} varies with V_T, over a frequency range of 5.000 to 5.500 MHz.

$$K_{VCO} = 2\pi \frac{\Delta f_{VCO}}{\Delta V_T}$$

$$K_{VCO}(5.0\,MHz) = 2\pi \frac{(5\,MHz - 4.9\,MHz)}{(1.75 - 1.25)}$$

$$\approx 1.257\,(10^6)\ rad/sec/volt$$

$$K_{VCO}(5.5\,MHz) \approx 2\pi \frac{(5.6\,MHz - 5.5\,MHz)}{(9.30 - 7.00)}$$

$$\approx 0.2732\,(10^6)\ rad/sec/volt$$

programmable divider

The programmable divider is a digital circuit that provides simple programmable integer frequency division. Frequency division is possible to beyond 1 GHz using available ICs.

The gain of the programmable divider is:

$$K_N = \frac{1}{N}$$

Since K_N is a function of N, two values of K_N must be specified. For example, a synthesizer is required to tune 5.00-5.50 MHz in 10 kHz steps. If a simple loop is designed as in **fig. 2**, then the reference frequency will be 10 kHz and:

$$N(5\,MHz) = \frac{5\,MHz}{10\,kHz} = 500$$

and $K_N(5\,MHz) = \frac{1}{N} = 0.002$

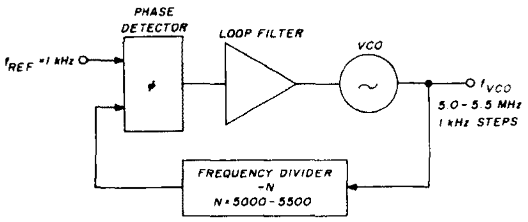

fig. 3. A 5.000-5.500 MHz synthesizer that tunes in 1 kHz increments.

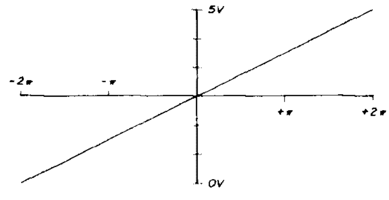

fig. 4. CD4046 phase detector transfer function.

$$N(5.5\,MHz) = \frac{5.5\,MHz}{10\,kHz} = 550$$

and $K_N(5.5\,MHz) = \frac{1}{N} \approx 0.00182$

Programmable dividers are often programmed from microprocessors, thumbwheel switches, or digital counters which may also drive a display.

loop filter

The loop filter is always the last circuit to be designed because it requires values of K_ϕ, K_{VCO}, and K_N to determine component values. The other information needed is ω_β, or loop bandwidth, and ξ, or damping factor. As a rule

$\omega_\beta \leq \frac{2\pi f_{REF}}{100}$ and $0.7 < \xi < 5$ with $\xi = 1$

a commonly chosen value.

The loop filter provides two important functions – it filters the error voltage from the phase detector so that the VCO receives a "clean" tune voltage (any signal or noise on the VCO tune line would modulate it), and it controls the loop parameters. It controls loop stability, determines lock time, and influences the synthesizer phase noise and spurious signal performance. Because of its important effect on loop performance considerable attention should be placed on careful filter design.

The active loop filter described in **fig. 6** provides better control over loop parameters than passive filters.

Component values for R_1, R_2, and C_1 are calculated as follows:

$$\omega_\beta \leq 2\pi \frac{f_{REF}}{100} \quad \textbf{(1)}$$

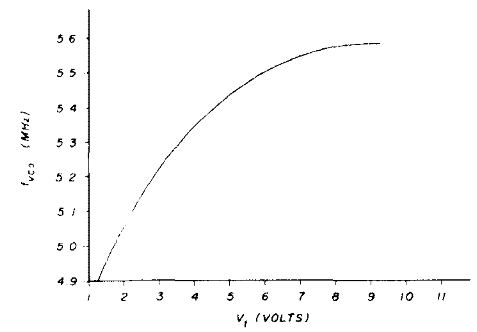

fig. 5. Example of f_{VCO} versus V_t showing nonlinear transfer function due to varactor diodes.

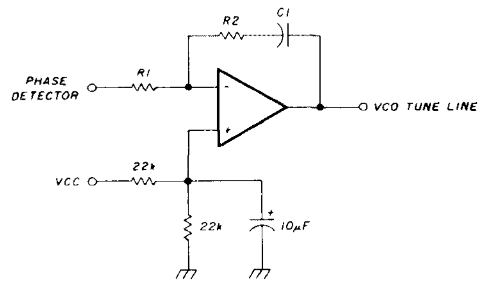

fig. 6. Op-Amp integrator used as active loop filter.

$$\omega_n = \frac{\omega_\beta}{\sqrt{2\xi^2 + 1 + \sqrt{(2\xi^2 + 1)^2 + 1}}} \quad \textbf{(2)}$$

$$t_1 = R_1C_1 = \frac{K_\phi K_{VCO} K_N}{(\omega_n)^2} \quad \textbf{(3)}$$

Where K_{VCO} and K_N are evaluated at the highest VCO frequency,

$$t_2 = R_2C_1 = \frac{2\xi}{\omega_n} \quad \textbf{(4)}$$

Choose a common value for C_1, such as 2.7 μF, and then calculate R_1 and R_2 to determine whether they come close to common component values.

$$R_1 = \frac{t_1}{C_1} \qquad R_2 = \frac{t_2}{C_1} \quad \textbf{(5)}$$

For example, a synthesizer required to tune from 5.000-5.500 MHz in 1 kHz increments might appear as in **fig. 7**.

A CD4046 used as the phase detector has a K_ϕ of 0.397 volts/radian. The VCO is the same one described earlier where $K_{VCO} = 1.257$ to 0.273×10^6 *radians/second/volt*.

With the loop locking at 5.000 MHz,

$$N = \frac{5\,MHz}{1\,kHz} = 5000 \text{ and}$$

$$K_N = \frac{1}{5000} = 200 \times 10^{-6}$$

With the loop locking at 5.500 MHz,

$$N = \frac{5.5\,MHz}{1\,kHz} = 5{,}500 \text{ and } K_N = \frac{1}{5500} = 181.82\,(10^{-6})$$

With a reference frequency of 1 kHz and $\xi = 1$, enough information is now available to design the loop filter.

Evaluating **eqs. 1-5** we obtain:

$$\omega_\beta \leq 2\pi \frac{1\ kHz}{100} \text{ or } \omega_\beta \leq 62.83\ rad/sec$$

$$\omega_n = \frac{62.83}{\sqrt{2(1)^2 + 1 + \sqrt{[2(1)^2 + 1]^2 + 1}}}$$

$$\approx 25.3\ rad/sec$$

$$t_1 = R_1C_1 = \frac{(0.398)(0.2732)(10^6)(181.82)(10^{-6})}{(25.3)^2}$$

$$\approx 30.89\ ms$$

$$t_2 = R_2C_1 = \frac{2(1)}{(25.3)} \approx 79.1\ ms$$

letting $C_1 = 2.7\ \mu F$ then:

$$R_1 = \frac{30.89\ ms}{2.7\ \mu F} = 11.4\ kilohms$$

$$\approx 12\ kilohms$$

$$R_2 = \frac{79.1\ ms}{2.7\ \mu F} = 29.29\ kilohms$$

$$\approx 33\ kilohms$$

If R_1 and R_2 had not come close to convenient values, then another value of C_1 would have been tried and R_1 and R_2 recalculated.

With loop filter components calculated, (see **fig. 8**) a quick check is required to ensure proper loop operation with the chosen filter component values at both extremes of the synthesizer frequency range.

At $f_{VCO} = 5.000\ MHz$

$$t_1 = R_1C_1 = (12\ kilohms)(2.7\ \mu F) = 32.4\ ms$$

$$t_2 = R_2C_1 = (33\ kilohms)(2.7\ \mu F) = 89.1\ ms$$

$$\omega_n = \sqrt{\frac{K_\phi K_{VCO} K_N}{t_1}}$$

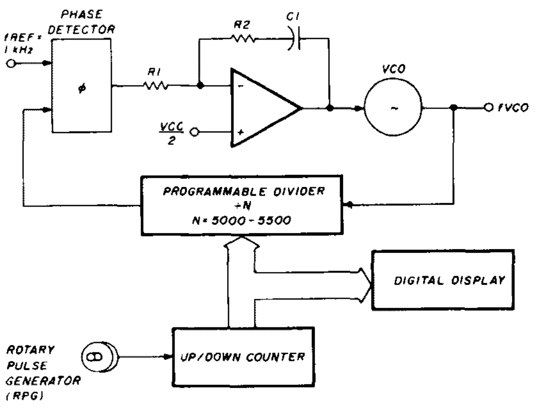

fig. 7. Synthesized VFO example loop.

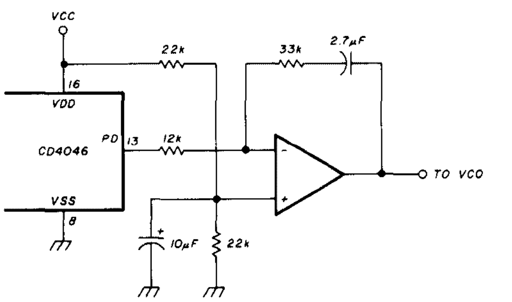

fig. 8. Simple Op-Amp integrator loop filter for example loop.

$$= \sqrt{\frac{(0.398)[(1.257)(10^6)][(200)(10^{-6})]}{32.4\ ms}}$$

$$= 55.57\ rad/sec$$

$$\xi = \frac{(\omega_n)\, t_2}{2} = \frac{(55.57)\ 89.1\ ms}{2} = 2.47$$

With ξ greater than 0.7 the loop is considered stable.

At $f_{VCO} = 5.500$ MHz

$$t_1 = 32.4\ ms$$

$$t_2 = 89.1\ ms$$

$$\omega_n = \sqrt{\frac{(0.398)\ [0.2732\ (10^6)]\ [181.82(10^{-6})]}{32.4\ ms}}$$

$$= 24.7\ rad/sec$$

$$\xi = \frac{24.7\ (89.1\ ms)}{2} = 1.1$$

Both ω_n and ξ are indeed very close to the design values; therefore, so will ω_β. The loop is definitely stable at both extremes of the synthesizer range and therefore at all points in between. This synthesizer should work fairly well.

conclusion

This first article demonstrates basic PLL synthesizer theory and design. Future articles will provide a more thorough explanation of the PLL and show more accurate methods of design. Methods to improve performance and test the loop will be included.

In the final article a 5.000-5.500 MHz synthesizer design will be presented. Performance of the completed synthesizer will then be compared with the initial design goals.

ham radio

Amateur packet radio: part 2

Understanding the TNC

This article describes the inner workings of a terminal node controller (TNC), with emphasis on those aspects which are novel and useful to others interested in implementing digital radio systems. The discussion is based on the TNC designed by the Tucson Amateur Packet Radio Corporation (TAPR).[1]

As shown in **fig. 1**, the TNC is essentially a special-purpose microcomputer. In many ways it is very much the same as any small computer system in that it contains a central processor, memory, and input/output (I/O) sections. The TNC differs from the average home computer in its I/O design, however, and we shall focus on these features.

The TAPR TNC uses a 6809 microprocessor, with 24K bytes of read-only memory (ROM) for program storage, and 6K bytes of read-write memory (RAM), for message buffers and other temporary data. The serial I/O port conforms to the EIA RS-232-C specification and is used to communicate through a terminal or with a computer. A dual 8-bit parallel I/O port is available for auxiliary use. A crystal-controlled clock provides system timing for various parts of the TNC.

The components of the TNC that make it a packet radio controller, and that could be added to a personal computer for a home-brew system, are the HDLC controller and the modulator/demodulator (modem). The HDLC controller is an LSI circuit which provides a convenient means for implementing much of the level 1 and level 2 protocol discussed in part one of this series (*ham radio*, July, 1983). It acts as a bidirectional digital port between the computer and the modem.

Equivalent to an RTTY terminal unit, the modem is a key part of the TNC, and contains the interface circuit that ties the computer to the station radio. It generates tones whose level can be adjusted for compatibility with the radio used, and its audio can be keyed to generate a Morse code station identification. The circuitry provides transmitter PTT line keying and a fail-safe timer to prevent excessively long key-down. The demodulator can be easily configured to accept audio from different radios, and includes LED level indicators for adjusting the receiver volume.

The versatile-interface adapter (VIA) block includes two 8-bit parallel I/O ports which communicate with the nonvolatile (RAM) semi-permanent storage, read user-settable switches, and control the modem. It also includes two counter-timers which provide interrupts for software timing and a programmable clock signal for the HDLC controller. The nonvolatile RAM, a Xicor NOVRAM™ which stores 32 bytes of information without battery or other standby power, represents a new technological achievement that should have wide application in Amateur Radio.[2]

modem and radio interface

Fig. 2 shows the radio and audio interface circuitry. Receiver audio output is buffered by one-half of U6 dual op amp. Provision is made at jumper JP9 for attaching a load resistor in place of the speaker. The audio passes through an LED limiter/level-indicator

The TAPR TNC will soon be available in the form of bare boards, documentation, and partial or complete parts kits. Please send an SASE to TAPR, P.O. Box 22888, Tucson, Arizona 85734, for details on price and availability, as well as further information on packet radio.

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to the MF-10 dual switched-capacitor audio input filter, which is configured as a highpass section followed by a lowpass section. Both sections operate in Mode 3, as described by the manufacturer,[3] with a clock input of 115.2 kHz produced by the TNC system clock. The response of the filter is optimized for typical transceiver combinations using a computer-aided design procedure.[4]

The necessity for this filter is dictated by the audio spectrum of 1200 Hz and 2200 Hz NRZI data at 1200 baud. This spectrum, shown in **fig. 3**, suggests that ideally the system should exhibit flat response from below 500 Hz to above 2900 Hz. In fact, the typical overall response measured using a pair of 2-meter fm transceivers is shown in **fig. 4A**. Without proper filtering the rolloff shown prevents data from being demodulated much above 600 baud. The filter with the eight programming resistors shown (slightly different from those on the TAPR Beta TNC) restores the response to that shown in **fig. 4B**, and seems a good compromise for a wide variety of fm rigs.

The filtered audio is demodulated by the XR2211, which is configured as recommended by Exar[5] for demodulating 1200 Hz and 2200 Hz 1200 baud data, except for the lock-detect filter at pin 3. For better immunity to false lock indications this filter's time constant was increased. In addition to digital data from pin 7, the lock detect signal at pin 5 is required by the software to monitor channel activity.

The MF-10/XR2211 demodulator combination works well with a wide variety of fm transceivers. However, the lowest bit error rates for a given degree of receiver quieting will be achieved only by custom tailoring the input filter to produce the proper response for a specific transceiver pair. Normal experience is that data will be received perfectly under "full quieting" conditions, but deteriorates rapidly as the noise level goes up.

Data originating on the TNC at the TxD output of the WD-1933 generates phase-continuous AFSK via the XR2206 modulator. As in the case of the demodulator, Exar's recommended values were used for loop components.[5] A control signal generated by the TNC's VIA under software control is used at point E to key the AFSK signal on and off. This permits the software to generate the CW identification and to eliminate modulator output except when actually sending packets. The modulator output is buffered by the second section of U6 before going to the microphone input of the transmitter.

The remaining circuitry of **fig. 2** grounds the radio's PTT line to key the transmitter whenever the WD-1933 $\overline{\text{MSCOT}}$ output is brought low. To prevent channel lockup a NE555 one-shot "watchdog" times

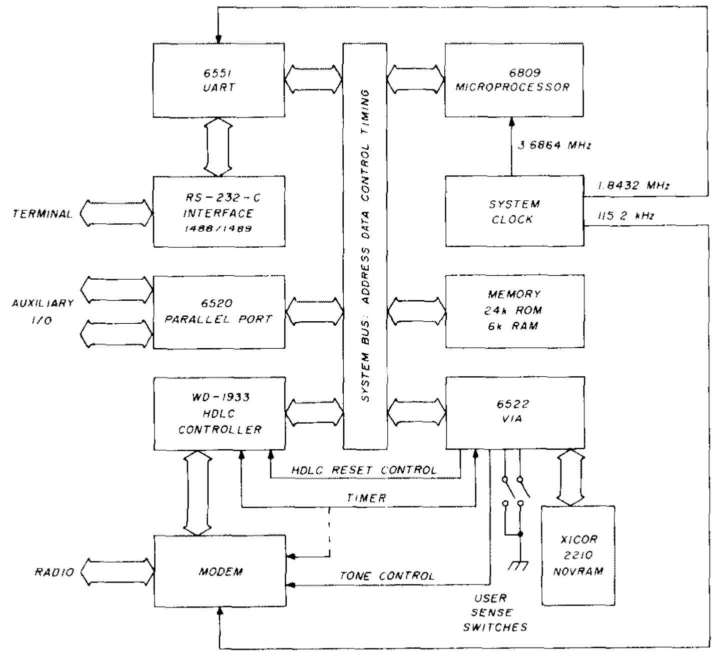


fig. 1. Block diagram of the TAPR Beta TNC, showing system architecture.

out after approximately 14 seconds, and $\overline{MSCOT}$ must be toggled high to restore PTT operation. This simple circuit has proven invaluable.

special digital hardware

The TAPR TNC includes digital circuitry that sets it apart from ordinary personal computers. Some de-

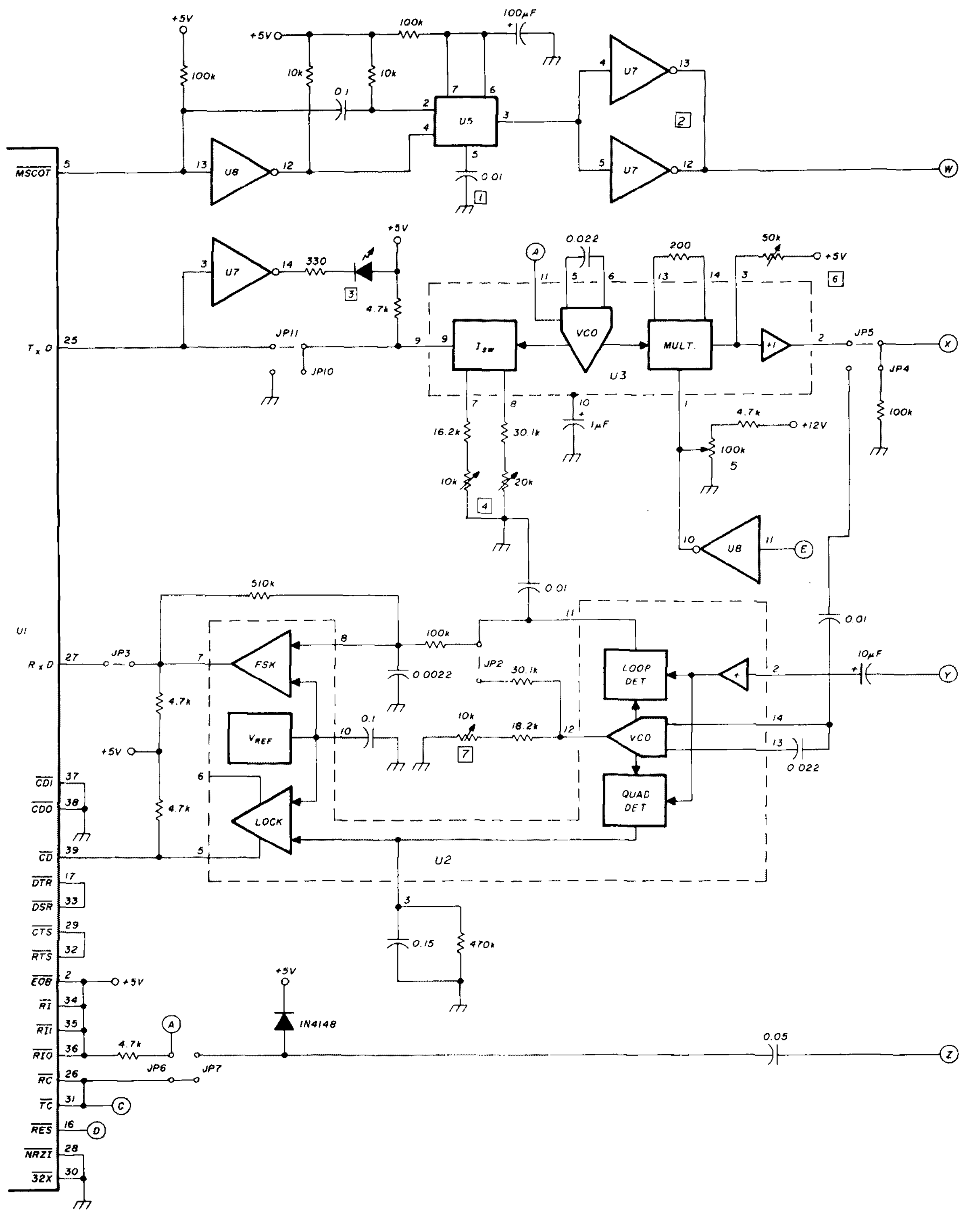
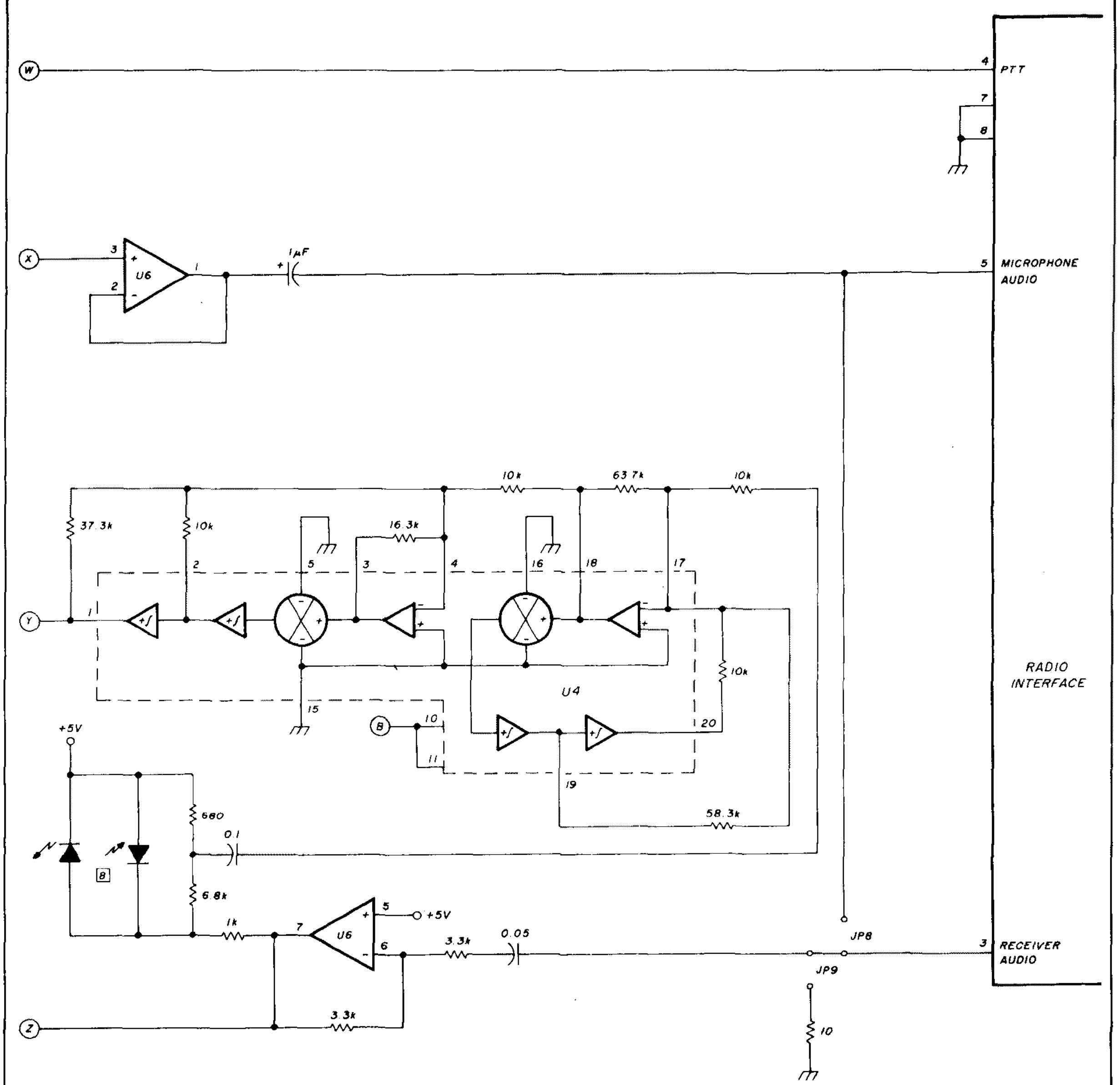


fig. 2. Schematic of the radio interface portion of the TNC, showing the modem components, input filter, and PTT circuits.

tails of these circuits will now be discussed; if you have a TNC, this information will help clarify some of its design considerations. Should you choose to homebrew a packet adapter for your computer, the discussion will serve as an example.

The most important such chip on the TNC is the HDLC controller. There are several HDLC controller chips on the market today, and more are being introduced regularly. This is fortunate for the would-be TNC designer, because the HDLC chip relieves him of a fairly complex hardware design (typically 19 SSI and MSI ICs) or an equivalently complex software



U1 WD-1933 HDLC controller
U2 XR2211 demodulator
U3 XR2206 modulator
U4 MF-10 switched-capacitor filter
U5 NE555 timer
U6 CA3240 dual op amp
U7 XR2203/ULN2003 Darlington driver array
U8 7406 open collector hex inverter

1 Approximately 15 second "watchdog" timer
2 PTT driver
3 Transmit data indicator
4 Adjustment for modulator frequencies
5 Adjustment for no-tone null
6 Modulator amplitude adjustment
7 VCO center frequency adjustment
8 Input level indicator-clipper

A Modulator oscillator connection to JP6
Connects to VIA timer input for calibration
B MF-10 clock input at 115.2 kHz from system clock
C VIA timer input/output
Operation: HDLC clock input from VIA timer output
Calibration: Modem signals to VIA timer input
D HDLC reset from VIA control output
E Modulator tone switch from VIA control output

fig. 2. cont.

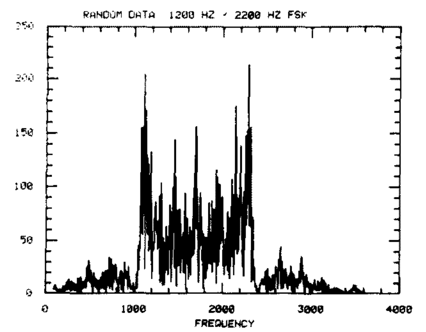


fig. 3. Sound spectrum of a one-second burst of 1200 Hz/2200 Hz 1200 baud random NRZI data. This data was generated and analyzed by computer.

program (in assembly language) to implement those parts of the HDLC protocol standard adopted by the overwhelming majority of packeteers.

As mentioned in the previous article, the most frequently used digital signalling technique employs Non-Return to Zero, Inverted (NRZI) encoding. This means that a digital zero is encoded as a transition from a high to a low or vice-versa, while a one is passed as the absence of a transition. The result of this, along with "bit-stuffing" (in which a zero is inserted by the sending station after five consecutive ones and stripped out at the receiving end), is that the clock signal can be recovered from the data stream. A phase-locked clock is necessary to ensure proper recovery of the data in this synchronous data mode, since the data must be latched in the middle of a bit and not, for example, just as a transition occurs.

Clock recovery is fairly straightforward using a phase-locked loop (PLL). Fortunately, the NRZI scheme is also widely used in the commercial world, so a few manufacturers have included a Digital PLL (DPLL) on their HDLC controller chips. In order to minimize the number of chips used in the TNC, both the VADCG and the TAPR designs incorporate these HDLC controllers.

Among single-channel devices, only the Intel 8273 and the Western Digital 193X series incorporate the DPLL, and the TAPR TNC uses the WD-1933, as it is generally about half the price of the Intel device. Of course, nothing is free, and some special considerations apply when interfacing this chip to a microcomputer. The software must take account of the inverted data bus of the WD-1933 which treats zeros as ones and vice-versa. In addition, the interrupt lines must be buffered and inverted prior to connection to the control bus. Furthermore, this chip requires a baud rate clock 32 times the data rate (for 1200 baud this means a 38.4 kHz clock) to drive the DPLL when using the NRZI mode, and also requires a special reset signal to be applied after the baud rate clock has been applied.

In exchange for these interfacing considerations, the HDLC controller provides automatic generation of pre-frame and post-frame flag bytes for synchronization, transparent bit-stuffing on transmit and unstuffing on receive, recovery of the clock signal from the incoming data stream, calculation of the Frame Check Sequence (FCS) used to validate data integrity on both transmit and receive, and automatic detection and reporting of errors in sending or receiving a frame. All in all, the usefulness of these LSI devices more than compensates for any interface difficulties.

In order to supply the HDLC controller with the needed reset and clock signals, and to provide other services, the TAPR TNC incorporates a 6522 Versatile Interface Adapter (VIA). This unit contains a pair of 8-bit parallel ports, which can be set on a bit-by-bit basis for input or output. Two of the four handshaking lines provided are used as single-bit control outputs. A pair of 16-bit counter-timers are also provided.

One of the control lines is used to provide a soft-

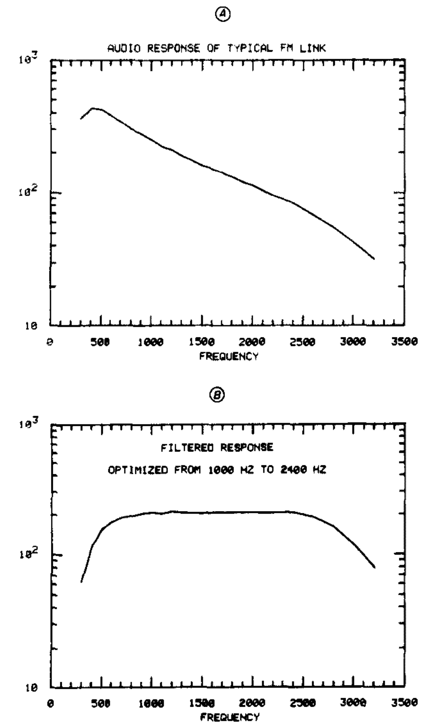


fig. 4A. Measured frequency response using a pair of 2-meter FM transceivers. fig. 4B. The same frequency response after filtering. The filter design was optimized for the range 1000 Hz to 2400 Hz.

ware controlled reset to the HDLC chip, while the other is used to effect a tone on-off command to the modem. This allows generation of an easily-copied CW station identification, as well as enabling an operator to insert a voice signal over the channel without disconnecting the TNC from the radio.

Two lines of one of the 8-bit ports (port B) connect to the internal 16-bit counter-timers. One timer is used as a software-controllable baud rate generator for the HDLC chip. This not only allows the operator a simple means for control of the baud rate, it also allows generation of non-standard baud rates, such as the 400 baud used in the AMSAT Phase III satellite.

The other timer is used for calibrating the modem frequencies and for primary system timing. From this clock are derived all the various clocks that must be updated for proper operation of the packet station of which the TNC is an integral part.

Two lines on the VIA are used to test the settings of a pair of switches on the board. These switches may thus be read by the software and are presently being used to tell the TNC whether to use the default parameter settings found in the system EPROM or whether to take these parameters from NOVRAM.™

The remaining lines from the VIA parallel port are used for the NOVRAM™ interface. This helps prevent accidental alteration of NOVRAM™ parameters, as well as easing system bus timing constraints. The NOVRAM™ is a nibble-oriented device, meaning that its data bus is only 4 bits wide, rather than the 8-bit bus width of the host microprocessor. It also has six address lines and four control inputs. The control lines allow for device selection, read/write control of data between the RAM portion of the chip and the data bus, recall of the contents of the Electrically Eraseable Programmable Read Only Memory (EEPROM), and storing of data from RAM into EEPROM.[2]

The presence of the NOVRAM™ permits long-term storage of parameters peculiar to the station, such as the call sign and terminal characteristics. In addition, infrequently adjusted parameters, such as those associated with the timing of data retries and other link activity, may be stored. Without such a long-term storage function there are only two choices. Either the operator must enter all necessary information every time the unit is powered up (which is not too practical), or the various parameters must be "burned-in" at assembly time, meaning that the operator must have his EPROMS erased and re-programmed every time he wants to change baud rates, call sign, station ID, and so forth.

controller software

The software present on the TAPR TNC is organized on two levels. The High Level Routines (HLRs) implement the machine-independent logical processes associated with protocol decisions and response to user commands. These routines know nothing of the hardware details of the TNC and, in fact, are written in a transportable high level language (Pascal). As the HLRs require data transfers or status information they call subroutines contained in the Low Level Routines package (LLRs), and leave the nitty-gritty details of interrupt service, terminal editing features, and timer maintenance to the LLRs. The LLRs, naturally, are written in 6809 assembly language, and are definitely not transportable. However, the logical organization of the LLRs is universal and should serve as a model for other implementations of packet radio.

The HLRs can be divided roughly into two major parts. One implements the command protocol, allowing the user to request connect and disconnect packets, control the digital-relay function of the station, and perform other tasks as necessary. For the TAPR TNC, this section consists of a command parser which compares a string of characters from the terminal with a list of commands and takes appropriate action when a match is found. In addition to issuing connect and disconnect packets, the user can alter program parameters, save the parameters in nonvolatile RAM, display current parameter values, identify in Morse code, change input mode, or enter a special service routine. For maximum flexibility in a test environment, the parser controls some sixty parameters, including the operator's call sign, terminal attributes, input editing features, radio interface characteristics, packet baud rate, and timing parameters[6].

The other part of the HLRs is a procedure which implements the packet radio protocol. This section assembles and disassembles packets, maintains information about the link status (for example, to whom you are connected), keeps track of unacknowledged outbound packets, acknowledges inbound packets, and sends supervisory packets as required by whatever protocol is implemented. This routine watches a clock to time retransmissions of unacknowledged packets, formulates input from the terminal into packets, and passes the contents of received packets to the terminal.

Both sections of the HLRs depend on the LLRs to maintain buffers and perform I/O under interrupt control. The LLRs also update clocks on receipt of timer interrupts and service the non-volatile RAM. When the program is required to transmit a message to the terminal or to send a packet, the information is actually loaded into a buffer to be sent when the peripheral component is ready. Similarly, the program reads incoming information not directly from

the peripheral, but from a buffer. The presence of complete messages to be read is signalled by flags set by the low-level input routines. This makes it relatively easy to implement the protocol in a high-level programming language without direct access to the peripheral devices.[7]

The utility of the HLR/LLR separation cannot be overemphasized. It allowed, for example, the two sections of the HLRs (command parser and packet protocol generator) as well as the LLRs to be developed independently by three people, in two different cities, on three different computers, prior to the final integration onto the TNC.

program structure

The structure of the packet protocol section of the HLR is shown in **fig. 5**. This procedure is part of an infinite loop in which all routines alternately check for tasks to be done.

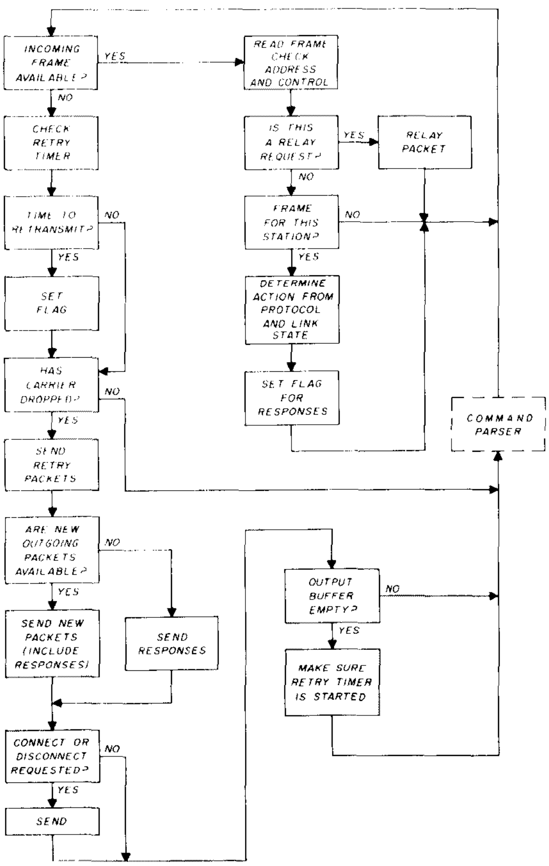


fig. 5. Flowchart for operation of the packet protocol HLR routine.

The first half of the procedure is concerned with reading incoming packets and determining the appropriate action to take. The action taken on receipt of a packet addressed to this station is determined by the protocol.[1,8] Several possible *link states* are defined, which are stages in the communication sequence starting with a connect request and ending with a disconnect request. For each type of packet received (specified by the CONTROL field) there is a prescribed action depending on the link state. If the action involves sending a packet — say, an acknowledgment — a flag is set for the second half of the procedure. When all incoming packets have been read, the clock is checked to see if packets have been sent which should have been acknowledged by now.

The second half of the packet protocol procedure, which sends outgoing packets, is entered only if the frequency is clear, indicating that all packets of a group have been received. This is determined by monitoring the demodulator carrier detect signal. Outgoing packets are formulated with header information and moved to the outgoing packet buffer following any packets being retried. Acknowledgments are sent as part of the control information with these packets if possible; otherwise, a special acknowledgment packet is sent. Finally, any special supervisory packets requested either in the first section of the procedure or by the user are sent. When transmission is complete, the clock is started for packets which should be acknowledged.

I/O management

The interrupt-driven I/O routines contained in the LLRs basically form a simple operating system for supporting the HLRs.[9] In order to isolate the main program from the details of the hardware, all input and output is done through buffers. Since the HLRs do not examine incoming data until an entire line or packet has been received, terminal support such as character echoing, line-feed insertion, and response to character, line, and packet delete instructions (implemented by single editing characters) are managed by the low-level interrupt routines.

The structure of a typical buffer is shown in **fig. 6**. There are four buffers, input and output buffers for terminal and radio data, each of which is accessed by an insertion pointer and a removal pointer. An input buffer, for example, has an insertion pointer which is advanced by an LLR interrupt routine as data is read from a peripheral device, and a removal pointer which is advanced by the HLR as it reads the data. All buffers are circular, meaning that when a pointer reaches the top of the buffer space it is moved back to the bottom. Input buffers require additional pointers to mark the beginning of a string which may be deleted by an editing command from the terminal, or

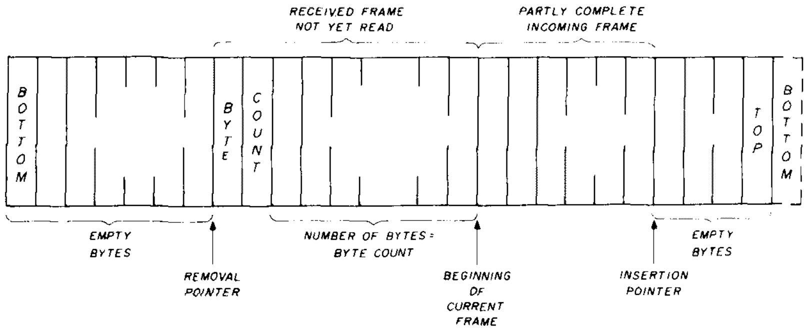


fig. 6. A typical data buffer, showing pointers, and order of placement of data. The buffer shown is the packet input buffer, and is similar to the other buffers.

in the case of the packet input buffer, by an error occurring during receipt of a packet. Since a data string can be any length, the end of a packet or command line must be marked, either by a special character in the buffer or by a byte count at the beginning of the string.

interrupt handling

Only one hardware interrupt-request input of the 6809 microprocessor is used in the TAPR design — all interrupt lines are wire-ORed together. This means that when an interrupt occurs, each peripheral which could have generated it must be queried in turn, and an appropriate routine selected from a dispatch table when the cause of the interrupt is identified. Since more than one device could be in need of service at once, the order in which the devices are queried determines the interrupt priorities, which are as follows:

1. UART (terminal) input
2. UART output
3. VIA timer interrupt
4. HDLC (radio interface)
5. Parallel port input
6. Parallel port output

The serial input port is given highest priority, since if a character is not read before a new one is received, it is lost. The radio I/O interrupts are placed relatively low, since servicing the WD-1933 chip is complex and potentially time-consuming. Data lost in either direction due to slow service of this chip will be detected as an error, and the packet will be retransmitted. If the parallel port is used for user I/O, it should be serviced last, since full "handshaking" is used, and a sending device will not send new data until the old data has been read.

The timer interrupt is generated as the timer counts down past zero. By examining the count, the service routine can determine the actual elapsed time and compensate for any delay caused by conflict with other interrupts. For this reason, the priority for servicing the timer interrupt is arbitrary. Compensation must be made for the fact that the two count bytes are read at different times.

The timer interrupt service routine has a special function. After the software counters have been updated, a general housekeeping routine is entered. Time elapsed since a carrier drop is monitored and a packet transmission may be started from this routine. The CW station identification is also sent at appropriate intervals, and the timer routine toggles the audio signal on and off to produce dits and dahs.

The WD-1933 HDLC controller generates interrupts for the following seven conditions:

Receive Interrupts (by priority)
- Data received
- Received message without errors
- Received message with errors
- Change in carrier detect state

Transmit Interrupts (no priority)
- Data requested
- Transmitted message without errors
- Transmitted message with errors
  (Abort signal sent automatically)

Since they may potentially be present in any combination, and querying the chip resets most conditions, all conditions must be checked on each interrupt. The only difficulty results when logically inconsistent or out-of-place interrupts occur. For example, the presence of both "Received message without errors" and "Received message with errors," or a carrier-detect change while the transmitter is keyed, may occur. This is solved by ordering the receive interrupt

priority as shown. Carrier detect can be ignored during transmission.

Transmit interrupts present a different sort of problem. The WD-1933 transmits HDLC frames automatically, but it must be commanded to send each section of the frame — flags, data, and frame-check sequence. While the transmit function is active, it generates regular interrupts. These are "Data request" if the data function is commanded, and otherwise "Transmit end of message." These interrupts are treated as equivalent, and the interrupt service function is determined by the progress of the packet being transmitted.

conclusion

In the first article of this series we described packet radio and the protocols in general use. In this article we have presented some details of the actual implementation of these concepts.

The TNC design presented represents the culmination of nearly two years of intensive effort by several Amateurs. These efforts resulted in both the formation of Tucson Amateur Packet Radio, a nonprofit R&D corporation of over 300 members worldwide, and the design and distribution of the TAPR TNC.

The TNC design was subjected to a Beta test with 172 boards placed at 19 sites. This test served to provide many useful improvements. Perhaps most importantly, it exposed literally thousands of Amateurs to this exciting new mode. We expect that soon there will be a rapid expansion in the use of this mode among Amateurs, and hope that you will join it.

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*Available from ARRL, Newington, Connecticut 06111 for $9.

ham radio

Measure characteristic impedance, balance, coupling coefficient, and saturation level accurately

testing baluns

In recent years baluns have become widely used in the antenna systems of most Amateur stations. Because of their popularity, many companies manufacture and advertise baluns regularly; their relative simplicity encourages many Amateurs to wind their own. Unfortunately, very little has been published about the performance requirements of baluns or even about which performance parameters are important. The performance characteristics of baluns *can* be measured, however, and by testing baluns according to the procedures described, users can learn what to expect from the baluns they install in their own antenna systems.

The tests described apply primarily to the familiar 1:1 transmission line balun of either the toroidal or ferrite rod type (see **fig. 1**), but may also be coincidentally appropriate for the 4:1 auto-transformer type balun.

Generally more than one test can be used to measure a given parameter. Most tests can be performed using equipment available to Amateurs. The choice of test will depend, to some extent, on whether the balun is purchased or homemade, since this will determine what terminals are available for testing. Because some tests require that the tertiary winding be disconnected from the top section of the main winding, not all tests can be applied to a purchased balun. All commercial baluns are factory-sealed and cannot be opened without breaking their cases; consequently, it is impractical to open the tertiary junction. With a homemade balun, all tests can be made before the tertiary winding is connected to the circuit.

balun operation

A balun serves two principal purposes. First, it provides two equal and opposite voltages to a balanced load with respect to ground. Second, the balun provides isolation between the balanced load (usually a dipole antenna) and an unbalanced transmission line (coaxial cable). Of particular importance is isolation between the coaxial outer conductor and the half of the dipole connected to the outer conductor. If this isolation is not adequate, then this half of the antenna essentially extends down the outside of the coaxial outer conductor and into the ham shack. This, of course, is undesirable. **Fig. 2** shows the problem graphically.

Your transmitter is actually a generator that drives a coaxial transmission line which is connected to a dipole. Assume the polarity is such that a current, I_1, flows into the center conductor from the left half of the dipole; an equal and opposite current, I_2, flows up the inside of the outer conductor. At the junction point between the outer conductor and the right half of the dipole, current I_2 divides.

Because of skin effect at radio frequencies, the inside and outside of the outer coaxial conductor may be thought of as two separate conductors. The division of the current I_2 into I_3 and I_4 depends on the relative impedances of the right half of the dipole and the impedance of the path down the coaxial outer conductor into your ham shack and through the power wiring to ground. If this length is an odd number of half wavelengths, the impedance will be low compared to the impedance of one-half a dipole (usually taken to be about 35 ohms). Much of the current I_2 will flow back down the outside of the coaxial and I_4 will be relatively high. Consequently I_3 will be low and different from I_1. In addition to causing the antenna to be fed asymmetrically, the outside of the coaxial can also be "hot" inside the shack, which not only creates operational problems, but introduces a safety hazard as well. However, if the length down the outside of the coaxial to ground is

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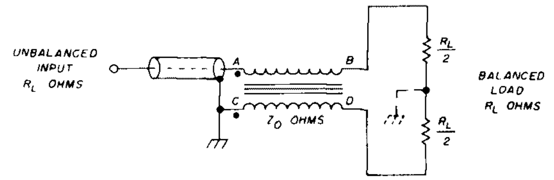

fig. 1. Typical 1:1, two-winding, transmission line balun. The characteristic impedance, Z_0, of the winding should be as close as practical to R_L. This balun will provide isolation, but not a balanced output unless the center-tap of the balanced load is grounded.

an odd number of 1/4-wavelengths, this impedance will be relatively high, forcing I_4 to be small in comparison to I_3 and the balancing dipole. The balun provides isolation between the right half of the dipole and the outside of the coaxial, and at the same time provides equal and opposite voltages and currents to the two halves of the dipole. Let's see how a 1:1 transmission line balun accomplishes these objectives.

1:1 balun

The simplest form of the 1:1 transmission line balun is shown schematically in **fig. 1**. Here a transmission line with a characteristic impedance (Z_0) is wound on a rod or toroidal ferrite core. One end of the winding is connected to the coaxial cable; the other end is connected to the balanced load, as shown. The characteristic impedance of the winding should be as close to that of the coaxial line and the load resistance as practical. One popular balun design[1] uses small 50-ohm coaxial (RG-141) wrapped around a 2-1/2 inch diameter (6.25 cm) toroid.

A commonly held view is that the inductance of the winding provides isolation and guarantees balanced voltages across the balanced load. Unfortunately, both these conditions may not occur in this design. The generator currents (I_1 and I_2 in **fig. 2**) are equal and opposite; hence there is no magnetic flux developed in the toroid and the inductance of the balun is essentially zero. The balun merely acts like an extra length of transmission line.

Assume an unbalanced condition, with current I_4 (**fig. 2**) flowing. Since there is no equal and opposite current flow, a magnetic flux will develop. If the inductance of the balun is sufficient, then the resulting counter EMF will limit I_4 and effectively isolate the balanced and unbalanced sides of the balun. The balun will therefore provide isolation. Counter EMFs will also develop on the inside of the coaxial shield and on the coaxial inner conductor; these will be in series opposition and do not affect the balun's operation.

Unless the center-tap of the balanced load is grounded or a tertiary winding is used, there is no ground reference point on the balanced side and no guarantee that the balanced side output is actually balanced with respect to ground. The degree of balance, if any, depends on parasitic inductances and capacitances and is not under control of the user. The only way the user can guarantee a balanced output is to actually ground the center-tap of the load. The lack of balance on a two-winding balun has been verified by actual measurements.[2]

To guarantee balanced output voltages and adequate isolation, it is necessary to provide a path for magnetizing current. Ruthroff[3] has stated that with the balanced load disconnected, there must be dc continuity between the unbalanced input and ground. The two-winding balun does not provide this continuity.

In order to guarantee balanced output voltages as well as provide adequate isolation, a tertiary winding, EF (see **fig. 3**), must be added. Note that the polarity of the tertiary winding is reversed.

If the voltage at the unbalanced input is V volts, the voltage at point B is V/2 volts since point B is halfway down the winding AB-FE with V/2 volts being developed in each winding. This is better shown when **fig. 3** is redrawn as an auto-transformer

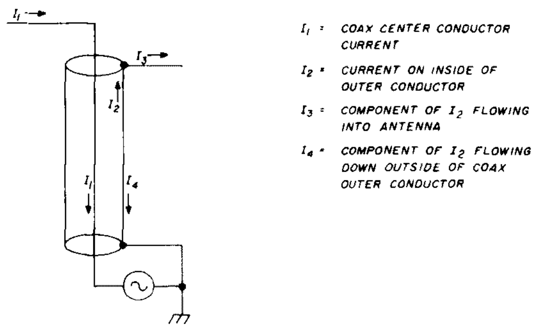

fig. 2. Lack of adequate isolation between antenna and feedline causes I_2 to divide into $I_3 + I_4$.

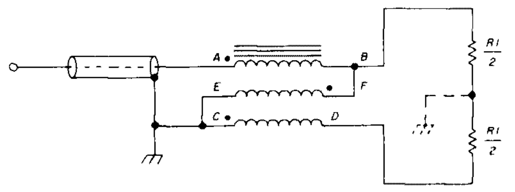

fig. 3. A 1:1 transmission line balun with a tertiary winding E-F. This design provides both isolation and balanced output voltages.

as shown in **fig. 4**. This arrangement guarantees that the balanced output voltages are balanced with respect to ground, provided that the coupling between tertiary and main windings is "tight."

Though **fig. 4** is drawn as an auto-transformer, the balun is nevertheless a transmission line device. Signal currents flow only through the transmission line windings and the input impedance/load impedance relationship follows the transmission line equation and not the auto-transformer law. With this thought in mind we will move on to the actual tests.

dc ohmmeter test

One of the simpler tests, to determine if the balun has a tertiary winding or not, is one that should be performed on any purchased balun. That test is important because a tertiary winding is absolutely essential if the balun is to work properly with most antennas.

This test consists simply of measuring the dc resistance between the unbalanced input terminals and ground with the balanced terminals open-circuited (see **fig. 5**). If a tertiary winding is present, this resistance should be a few tenths of an ohm and will appear on most ohmmeters as a short-circuit. An open circuit reading indicates no tertiary winding.

Using an accurate ohmmeter or Wheatstone bridge can provide other information. With the unbalanced side open-circuited, measure the dc resistance between each balanced load terminal and the grounded side of the unbalanced terminal. Each of these resistances should be one-half the value obtained in the first test. The success of this test ensures that each of the windings is the same length and that the balun is reasonably well balanced, at least in regard to dc.

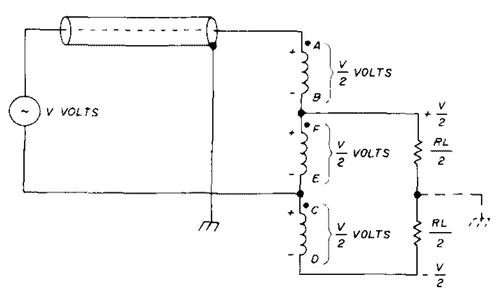

fig. 4. A 3-winding transmission line balun drawn as an auto-transformer. Though shown as an autotransformer it is still a transmission line device. The input versus load impedance relationship follows the transmission line equation and not the transformer law.

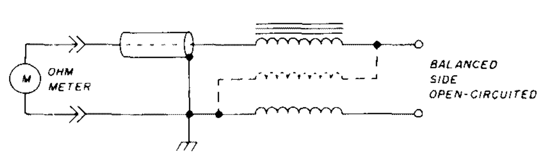

fig. 5. Ohmmeter test to determine if balun has a tertiary winding.

characteristic impedance

One of the most important parameters of any transmission line balun is its characteristic impedance which should be the same as the characteristic impedance of the transmission line with which the balun will be used. If too great a difference between these impedances exists, use of the balun may *cause* more problems that it cures.

There are several ways of measuring the characteristic impedance of a balun. The method used will depend on the measuring instrument available and on whether or not the balun is store bought or homemade.

Perhaps the most straightforward method of measurement is to take advantage of the fact that the characteristic impedance can be found by taking the square root of the input impedance with the far end open-circuited and short-circuited or:

$$Z_0 = \sqrt{Z_{oc} Z_{sc}}$$

While this approach is theoretically straightforward, it presents instrumentation problems. At some frequencies, the input impedance of the line will have a very high or very low resistive and/or reactive component for either an open- or short-circuited condition at the far end with one or more of these components outside the range of the measuring instrument. If it is possible to find a frequency or a test instrument where both open- and short-circuit measurements can be made, this method provides a convenient way to determine the characteristic impedance of the balun.

It is important to note that this test can *not* be used with a tertiary winding connected. If the balun is homemade, make the test with the tertiary winding in place but not yet connected to the main windings. If the balun is commercially made, you may have to figure out a way to open the balun and disconnect the tertiary winding. If this is not practical, use a different measuring technique.

A second method which does not put such severe requirements on the test equipment but is more time consuming is to measure the input impedance of the

balun with an arbitrary load impedance as the electrical length of the line changes, or place the value of load resistance across the balanced load that you think the balun characteristic impedance is, or that you would like it to be. Then measure the input impedance of the balun across the frequency range and see how close your guess was. **Fig. 6** shows the input impedance of a transmission line balun with four different values of load resistance: 65, 76, 84, and 101 ohms.

Starting at the top, notice that the input impedance with the 101 ohm load rolls off at higher frequencies. At first glance this roll-off appears as normal high-frequency drop off. However, looking at the 65-ohm load line, we see a "roll up" in the input impedance as the frequency increases. These two input impedance characteristics suggest the impedance inverting effect of a 1/4-wave transmission line whose characteristic impedance, which is what we are trying to find, is between 65 and 101 ohms.

Looking between 65 and 101 ohms, we see that the 84 ohm line rolls off slightly while the 76 ohm load response is practically flat — only a very slight roll-up. This indicates that the characteristic impedance of the balun is just above 76 ohms. If these measurements were made beyond a 1/4-wavelength, the slope of the curves would reverse and the frequency for a 1/4-wavelength could be determined. My equipment does not go high enough in frequency to do this, however.

If a General Radio 821 Twin-Tee admittance bridge[4] or similar instrument is available, a third approach may be used that gives not only the characteristic impeda ce but the electrical length of the winding as well. If the physical length of the winding (in inches or meters) is known, the velocity coefficient of the winding can also be determined from this data.

This test is based on the fact that a short-circuited transmission line 1/8-wavelength long has an inductively reactive input impedance equal to the characteristic impedance of the line. Similarly, if the far end is open-circuited, the input impedance presents a capacitive reactance equal to the characteristic impedance. This can be seen by examining the transmission line equation for the short-circuited case, which is the simplest:

$$Z_{in} = Z_0 \frac{Z_r \cos X + jZ_0 \sin X}{Z_0 \cos X + jZ_r \sin X}$$

when $Z_r = 0$ (short circuit load)

then $Z_{in} = Z_0 \dfrac{jZ_0 \sin X}{Z_0 \cos X} = jZ_0 \tan X$

when $X = \lambda/8 \text{ or } 45°$,

then $(\text{since} \tan 45° = 1)\ Z_{in} = jZ_0$

The open-end and short-circuit values of reactance are plotted versus frequency on the same piece of graph paper; the reactance at which they intersect is the characteristic impedance of the balun. An example of this is shown in **fig. 7**. Also, the frequency of intersection is the frequency at which the balun is 1/8-wavelength long. The electrical length at any frequency can easily be determined from this.

If the physical length of the line is known, the velocity coefficient of the transmission line, k, can be determined by calculating the 1/8-wavelength of the intersection frequency in free space and dividing this into the measured length of the balun winding.

$$k = \frac{\text{measured length winding}}{\text{calculated length } \lambda/8 \text{ free space}}$$

Because this test cannot be used when the tertiary winding is connected, it can be used only on home-made baluns or commercial baluns where the tertiary winding can be opened.

The limitations on the test equipment are that the impedance measuring device must be capable of measuring impedances with high resistive components and measuring reactive components in the range of the expected characteristic impedance at the frequency where the balun is 1/8-wavelength long.

winding inductance

The balun winding inductance is important because it determines the frequency range over which the balun can be used and also determines balun isolation. In general, the winding reactance should be about five times the characteristic impedance for a general purpose balun. You may want to use a factor of ten times the characteristic impedance for a precision or an instrument balun, however.

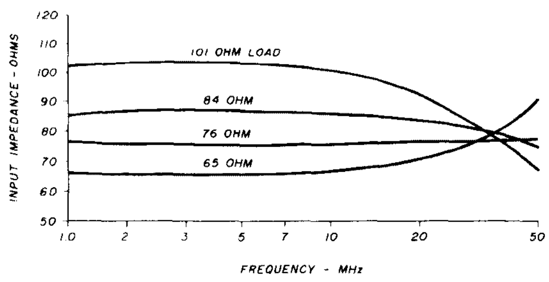

fig. 6. Input impedance versus frequency for 65, 76, 84, and 101 ohm load resistors.

As the frequency increases, the balun impedance increases until the inductance resonates with the stray capacity across the inductance. At this frequency, the impedance of the winding and the isolation are the highest. As the frequency is further increased, the impedance becomes capacitively reactive and decreases until series resonance occurs and the winding is effectively a short-circuit. The balun is obviously worthless at this frequency as the balun develops no isolation between the balanced and unbalanced sides. There is no problem in operating the balun through parallel resonance, but it should not be operated above the frequency where the impedance falls below about five times characteristic impedance. **Fig. 8** shows a typical inductance curve.

To perform this test, the tertiary winding must be disconnected so you may not be able to make these measurements on a commercial balun. The test arrangement is shown in **fig. 9**.

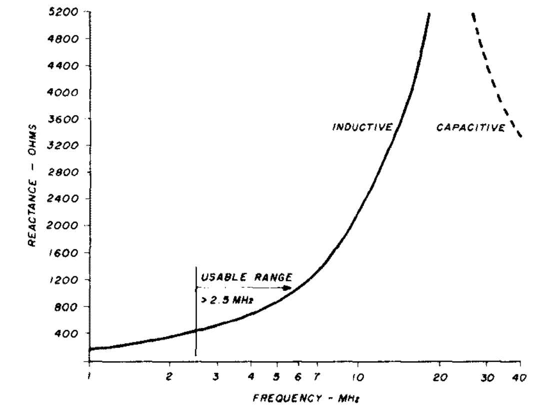

fig. 8. A typical balun reactance versus frequency characteristic. Isolation is at least five times the characteristic impedance from 2.5 MHz to above 40 MHz.

coefficient of coupling

The coefficient of coupling between the main winding and tertiary winding is important because it affects the degree of balance of the balanced output and also limits the high frequency response. To measure the coefficient of coupling, the tertiary winding must be disconnected from the main winding; again, this will restrict the testing to homemade baluns. The procedure is simple; measure the inductance of the main winding as described in the preceding test, with the tertiary winding open-circuited and again with the tertiary short-circuited. Use the equation:

$$k^2 = 1 - \frac{L_{sc}}{L_{oc}}$$

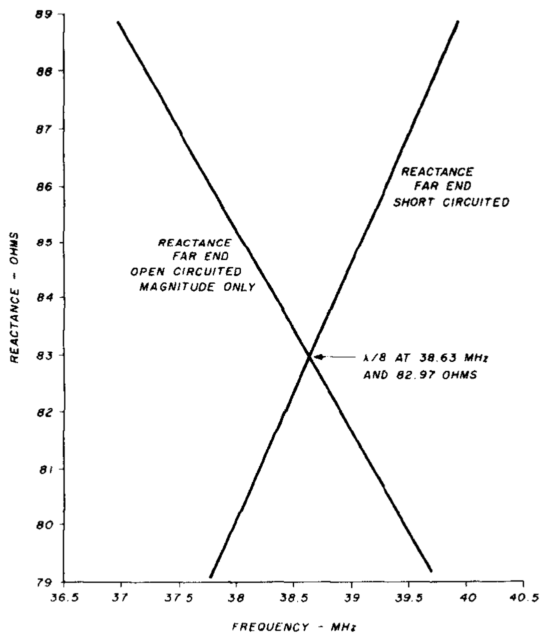

fig. 7. Open/short circuit curves of a balun transmission line showing a characteristic impedance of 82.97 ohms and a l/8 length at 38.63 MHz.

Values of k — not k^2 — should be at least 0.98 to 0.99. If the coefficient of coupling is less than about 0.98, you should expect problems, especially if broadband operation and/or a mismatch condition exists. Since this test involves measuring the inductance of the main winding, it is convenient to do it simultaneously with the inductance test.

achieving a balanced output

A very important performance characteristic of any balun is the degree of balance of the balanced output (assuming a balanced load). Fortunately, the test for this is easy to do, and a number of different approaches are possible.

The simplest and most direct approach is to measure the rf voltage between each side of the balanced load and ground over the frequency range of interest. If the input voltage is held constant, any unbalance or variations in the transmission through the balun will be apparent.

Another approach is to use a dual channel oscilloscope with one channel connected to each of the balanced terminals. This has the advantage that phase differences between the two halves of the balanced output can also be measured by the horizontal displacement of the two traces.

A variation on this test is convenient for measuring the electrical length of the winding; connect one scope channel to the balun input (unbalanced input) and the second channel to the high side of the balanced output of the balun. The balanced voltage to ground at this point should be one-half the input voltage for a 1:1 balun. You will probably want to synchronize the scope on the channel connected to the balun input. The electrical length of the winding can be determined from this measurement from the horizontal displacement of the two traces. Scopes with vertical channel responses of 30, 45, and even 60 MHz are now readily available to Amateurs, making this an attractive method.

The ideal method of measuring the balance is to use a Hewlett-Packard model 8405A vector voltmeter; this instrument measures the magnitude of two voltages and the phase angle between them. Unfortunately, this is a $5000 instrument and very few Amateurs can afford to spend this much for a voltmeter. If you are employed in electronics, see if your lab has one; it's a common instrument in rf labs. The vector voltmeter can also be used to determine the electrical length of the balun.

Another simple and useful test for estimating the degree of balance is to use a balanced load impedance composed of two resistors in series, with each resistor being one-half the value of the desired balanced load. The input impedance is then measured with the center-tap of these resistors both grounded and open-circuited. If the balun and load are well balanced, there will be no change in the input impedance of the balun when the center-tap of the load is grounded or ungrounded. By "grounding the center-tap," I mean connecting the center-tap to the grounded terminal of the unbalanced input. When I have performed this test on a well-designed balun, I have found that the change in input impedance is always less than the width of the calibration line of the dial.

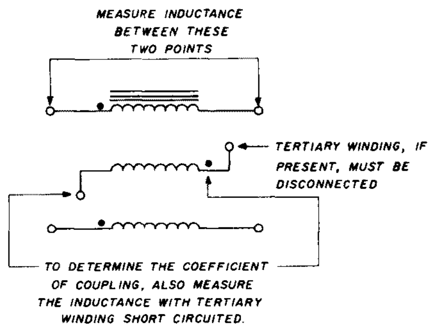

fig. 9. Circuit for measuring balun inductance and calculating the coefficient of coupling between main and tertiary windings.

As the balance test must be made with the balun in its final operating configuration, it can be made on commercial baluns as well as homemade ones. This is one of the simplest and most effective tests I am aware of for determining the effectiveness of a balun. As the balun test involves only measurements of input impedances, it is convenient to check balance when the input impedance is measured. To demonstrate the benefit of a tertiary winding, try making this test on a 1:1 balun without a tertiary winding.

short-circuit test

This test was first described by Reisert.[1] Readers have often called it to my attention after publication of my previous articles about baluns.[2,4,5] Basically, this test is intended to give an estimate of the isolation provided between the balanced and unbalanced sides of the balun.

This test is performed by measuring the input impedance at the unbalanced terminals with a normal balanced load connected to the balanced side. The input impedance is again measured when each of the balanced terminals is shorted to ground. If baluns provided perfect isolation, there would be no change in the measured input impedance; but because nothing is perfect, some change in input impedance should be expected. Despite extensive reading in the field, I have not yet discovered what constitutes an acceptable change in input impedance, but I would assume that a change of ten percent or less is acceptable. This would suggest that the series impedance of the balun is at approximately ten times its characteristic impedance.

However, this test must be approached with extreme caution. First, if the balun has a tertiary winding, shorting the high side of the balanced terminal to ground will also short-circuit the tertiary winding. As the tertiary winding is tightly coupled to the main windings, this will effectively short-circuit the main winding, thereby ruining the balun action. If the balanced-load center-tap is grounded, shorting either side to ground will also short-circuit one-half of the load resistance, which will obviously affect the input impedance.

My principal objection to this test, however, is that the test conditions alter the operating conditions of the balun. Grounding either balanced terminal provides a path for the magnetizing current (a dc path to ground) and also increases the voltage across the main winding by a factor of two — from V/2 volts to V volts where V is the unbalanced input voltage. For these reasons, I am not convinced that this test is really a reliable indication of balun performance.

core saturation

One final test that should be mentioned is magnetic saturation of the core. I have not tried this test myself, but it's easy enough to perform, at least in theory. Wrap three or four turns of insulated wire around the balun core and connect it to an oscilloscope. If the waveform on the scope is a sine wave, the core is not being saturated. The test must obviously be made at full power while connected to the actual load. This presents some practical as well as safety problems.

summary

I have briefly discussed the purposes and operation of a 1:1 transmission line balun and described seven tests that can be used to measure the characteristics of the balun. The tests include:

1. tertiary winding (using an ohmmeter)
2. determination of characteristic impedance
3. isolation determination by winding inductance
4. balance
5. coefficient of coupling of tertiary winding
6. electrical length
7. core magnetic saturation

The tests described above do not appear to require specialized test equipment or training and I feel that balun vendors should list the various technical parameters as do manufacturers of other products. This information would benefit the users because they would be better able to choose the balun that best met their requirements. Perhaps if balun manufacturers were to share their test results with consumers by including technical specifications in their promotional materials, users could be spared some of the time and effort testing requires. Armed with such information, users would be better able to choose the balun that best meets their needs.

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ham radio

ham radio TECHNIQUES

Bill Orr, W6SAI

Last winter and spring the West Coast experienced severe weather with flooding, land slides and heavy property damage. Radio Amateurs supplied emergency communications in many cases. One important point learned by all concerned with these emergencies is that disasters occur suddenly and unexpectedly. Advance preparation is absolutely essential. In general, communications from home stations were of little use; hand-helds and portable stations carried the larger portion of the communications burden.

During these emergencies, many emergency communication coordinators found that the common "rubber duck" antenna on the hand-held unit was not suitable for emergency use. A better antenna was needed, but it had to be inexpensive and also rugged. A successful emergency antenna had been developed in Arizona for the Scottsdale Amateur Radio Club and the Arizona Repeater Association, and that design has been copied for use by the Red Cross and other emergency communications organizations.

the 2-meter J-pole

As described by Jack Hanny, KB7CH, of Scottsdale, the emergency J-pole antenna is light enough to be rolled up and carried in a tool box or emergency kit.

The J-pole is made from a 55½-inch section of TV "ribbon" twin lead. A ¼ inch of insulation is removed at one end (**fig. 1A**) and the wires soldered together. 16 inches above the short, a piece of the ribbon line is notched out and one lead is cut open. A ¼ inch of wire is removed. The break is then taped or covered with heat-shrink tubing.

The next step is to measure 1½ inches from the shorted end of the ribbon line and then carefully trim away the insulation to expose the two wires. Be careful not to nick the wires. The feedline is attached at this point (**fig. 1B**). Solder the center conductor of a random length of RG-58/U coaxial cable to the long wire of the ribbon line and solder the braid of the coax to the short conductor.

Jack made his coaxial cable about 12 feet long and placed a matching plug for his hand-held unit at the free end of the line. He wrapped the short section of ribbon line to the coaxial cable with string and covered the joint with tape or heat-shrink tubing.

The last step is to punch a small hole in the insulating web at the opposite end of the ribbon line and tie a section of heavy string to the top end of the antenna. This makes it possible to support the antenna from the branch of a nearby tree. An extra length of RG-58/U cable can be made up with matching connectors to be used if the antenna is to be hung from a greater height.

more on the sloper

A lot of words have appeared about slopers during the past decade. There's no doubt that it works, but the theory behind this unusual antenna is obscure. In brief, the sloper is simply a ¼-wavelength (approximately) wire, fed at the top end supported by the station antenna tower or mast. The bottom end of the wire is anchored a few feet above ground. The coaxial center conductor feeds the wire at the top, with the shield of the line attached to the metal tower, which apparently works as a ground point.

Dick, WD4FAB, has broadbanded a sloper by increasing its effective diameter with a four-wire cage (**fig. 2**). The bottom ends of the four wires must be interconnected; if they are not, each of the individual wires will take on its independent characteristics, resulting in some unpleasant bumps in the SWR curve. The WD4FAB sloper is suspended at the 50-foot point on a 58-foot-high tower.

comparing an inverted-V with a 5-band trap vertical

Jim, KW2W, compared a 5-band trap vertical to an 80-meter inverted-V used with an antenna tuner for operation between 80 and 10 meters. Jim says he has a very good location for a vertical — on a sandy beach, only about two feet above the salt water level. The soil beneath the antenna is always moist with salt water. The 5-band vertical was mounted atop a 10-foot pipe driven

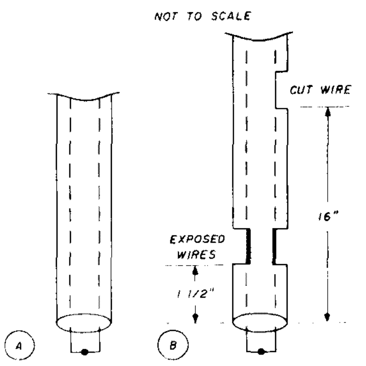

fig. 1. Construction details, 2-meter J-pole antenna.

into the soil and had two resonant radials for each band, for a total of 10 radial wires. The inverted-V was cut for the middle of the 80-meter band. The center of the antenna was about 50 feet high and the ends were about 15 feet high.

Over a period of time, Jim concluded that manmade and atmospheric noise was much less on the inverted-V and that reports received nearly always favored the inverted-V over the vertical.

Jim said, "I feel there's an antenna for every location — meaning that what works well in one place may work poorly someplace else. Antenna experimentation for a given location is worth the effort. The question of which antenna is better is really not applicable; the question of which antenna, *for a given location*, will perform better, is more precise."

the no-code license — a brief history

The present FCC proposal for a VHF/UHF "no-code" license brings back memories to old timers who remember the 1932 uproar over a similar suggestion.

1932 was a critical year for Amateur Radio. At the depths of the Great Depression, millions were out of work and industry was at a standstill. Many young men with plenty of time but little money turned to the fascinating hobby of shortwave listening. For a few dollars, or even less, an old battery-operated radio could be torn down and rebuilt into a simple shortwave receiver. Many newspapers carried columns about shortwave reception and shortwave clubs were founded for avid young listeners all over the country.

These enthusiastic SWLs soon discovered the Amateur bands, particularly the phone stations. Amplitude modulation was exclusively used for voice transmission in those days and the signals could be readily received on a one or two tube receiver.

A direct result of the SWL hobby was an expanded interest in Amateur Radio. Thousands of listeners yearned to be Amateurs and would have been — except for the bothersome task of learning the Morse code! Why was the knowledge of code required for "radiophone" operation, especially on the "ultra-high" frequencies above 10 meters?

The interest in a no-code license came to a head in May, 1932, when *Short Wave Craft* magazine, edited by Hugo Gernsback, announced the formation of the "Short Wave League" (**fig. 3**) devoted to "the Amateur who is not interested in code, but who is interested in the transmission of voice only." The Short Wave League had no dues or membership fees. The charter of the League was vague, but the editorial in the issue announcing the formation of the League was specific: the goal was to be the lifting of the code restriction on the Amateur "extra-short wavelengths." The May, 1932, editorial in *Short Wave Craft* promised that if "a sufficient number of letters were received, they would form a basis of negotiations between the League and the Federal Radio Commission."

In this manner the request for a no-code VHF license was created. Looking back, it seems unclear whether the Short Wave League was merely a gimmick to increase magazine circulation, or in fact represented an authentic desire for a no-code Amateur license. For a year or so *Short Wave Craft* was full of angry letters to

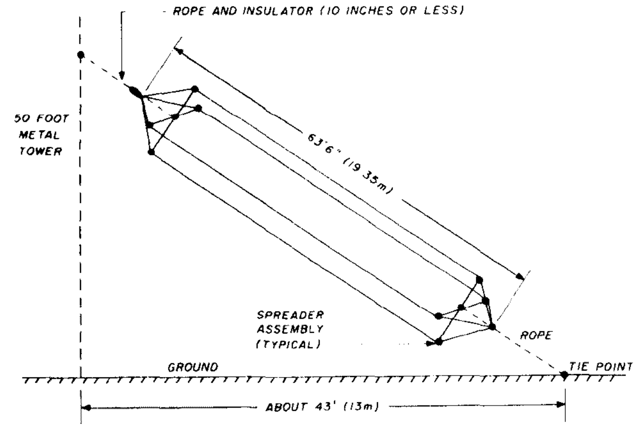

fig. 2. 80-meter WD4FAB broadband sloper. Spreader assemblies are made of 36-inch long fiberglass tubes. The four antenna wires are tied together about 24 inches past each spreader assembly. A nylon line may be required from the crossover point of the spreader to the outer insulator to prevent bowing. The SWR plot of the antenna falls below 1.4-to-1 at 3.5 MHz; 1.25-to-1 at 3.8 MHz and 1.4-to-1 at 4.0 MHz. Feedpoint resistance of the antenna may be changed by altering the slope angle. The resonant frequency is adjusted by changing length — no climbing necessary!

SHORT WAVE LEAGUE PLATFORM

1.—To help popularize short waves among the public.

2.—To assist authorities in the apprehension of criminals and in general crime prevention through the instrumentality of short waves.

3.—To demand enactment of new regulations enabling operators of phone transmitters to be licensed without code requirments, below 6 meters.

4.—To bring together short wave enthusiasts by promoting comradeship among them.

5.—To eliminate man-made static, the greatest enemy of short waves.

fig. 3. The battle for the no-code Amateur license was fought — and lost — by the "Short Wave League" in 1932. The platform of the League makes interesting reading today. Point 2, which seems unusual today, had meaning in 1932. Police radio was in its infancy and many local police organizations relied upon Radio Amateurs to relay messages for them. But the real purpose of the League was point 3, the no-code license for the "ultra-high" frequencies. As for point 5, we are still working on that problem today! (Material reprinted from *Short Wave Craft* magazine).

the editor expressing various views, pro and con, on the no-code proposal. It is interesting to note that arguments used then were strikingly similar to those used today. The principal difference between the situation in 1932 and that of today seems to be that the early no-code proposal was apparently a grassroots movement sponsored by the magazine and supported by many of its readers. The Federal Radio Commission (the predecessor of the FCC) had nothing to do with the launching of the proposal. Moreover, it seemed to have given little thought to it, since no official comment was made on the matter. Most Amateurs — and *QST* — ignored the uproar, hoping it would go away. And, sure enough, it did. The fad of shortwave listening died out quickly and by 1934, *Short Wave Craft* magazine turned pro-Amateur and the no-code proposal was forgotten for a few years.

The next no-code license uproar occurred shortly after World War II when the FCC proposed a "Citizen Radio Service" to be placed in the 400 MHz region. The proposal was greeted with little enthusiasm by hams and non-hams alike because no commercial equipment was available for the band. Furthermore, because of the very high frequency chosen, any homemade rigs were of the "squawk-box," super-regenerative type, which had poor sensitivity and selectivity.

Faced with a likely failure, the FCC next proposed to expand the unwanted VHF CB service into the so-called "industrial-scientific-medical" frequency range at 27 MHz, heretofore used by Radio Amateurs on a secondary basis. Radio Amateurs and others protested the plan, predicting that the FCC wouldn't have the manpower, desire or ability to police the operation properly. How true that prediction turned out to be! And as the MUF rose, providing long distance contacts at 27 MHz, CBers quickly discovered that with modified ham gear, linear amplifiers and big antennas, they could work worldwide DX just like real radio hams! And best of all, the FCC couldn't really catch them — especially if they had no license at all and their names weren't in the FCC computer! No-code license? Why worry about that when no license at all was necessary?

refighting the battle of 1932

Little was heard of the no-code proposal until a few years ago when it began to surface once again, possibly resurrected because of the monumental CB problem that arose about 1976. The CB channels exploded with activity after CB radio received national publicity during a truckers' strike. Unlicensed CB activity spread beyond the authorized channels, until today it occupies the spectrum from about 26 MHz to 27.99 MHz.

One solution to the CB problem was to give CBers another band. The 220 MHz ham band was proposed, but protests from Amateurs and the military finally defeated the idea. Probably without recalling the 1932 hassle, and with the hope of solving the CB problem, the FCC proposed a no-code license for Radio Amateur operation on certain VHF bands, reopening the old argument that had been settled decades ago.

Why has this idea resurfaced after 50 years? Is there a grassroots movement for a no-code license? Are the CBers enthusiastic about a no-code license? Are the Radio Amateurs enthusiastic about a no-code license? As far as I can determine, the answer to these questions is *no*.

If this conclusion is correct, who, then, wants a no-code license? (All eyes turn toward the FCC.)

The present problem, as I see it, is more fundamental than whether or not a no-code license structure is established. The root of the matter is *who will control the destiny of Amateur Radio in the United States?* Do Amateurs have a voice in their own destiny? A dangerous precedent can be set if the FCC ignores the feelings of the majority of Radio Amateurs and forces a new class of Amateur into existence, flaunting the time-honored foundation of Amateur Radio itself.

It would be easy to establish a no-code Amateur group. But, once created, it would be impossible to disband it. The speed at which the undertaking advances provides little time for reflection or judgment of the long-term possibilities.

The Morse code has been with us for a long, long time. It may be scoffed at by those who don't know it; on the other hand, it could be considered a badge of honor to those of us who use it and appreciate it. I think it would be a mistake of the first magnitude if a complete new class of Amateur licensee were to be artificially created who had no "feel" for the majestic scope of Amateur Radio — the Morse code included.

ham radio

antenna carriage
and track pole mount

Mounting a rotatable antenna on a utility pole can be easy, inexpensive

Some ingenious ways have been devised to raise antennas. These have included mounting a stationary mast on a rotatable base; digging a hole, setting a pole, and then cranking the mast up and down; raising and lowering a mast through the roof; making a tiltable mast, using gin poles; and using a mast with telescoping sections.

The technique I've devised consists of stringing two cables vertically on a utility pole, 10 inches apart. A pulley at the top of the pole serves as a sheave for the steel cable that raises and lowers the antenna-bearing carriage. The carriage rides up and down the vertical cables.

access to the top

Mounting the pulley requires the use of an extension ladder tall enough to reach the top of the pole. Start by mounting a curved block of wood on the top rung of the ladder as shown in **fig. 1**. This prevents the ladder from sliding sideways while positioned at the top of the pole. The base of the ladder should be secured by lashing it to stakes driven into the ground. Use guy ropes to keep the ladder from swaying.

pulley and cables

Because the top of the pole is 8-1/2 inches (21.6 cm) in diameter, a 9 inch (22.8 cm) die-cast aluminum pulley is required. The pulley is attached to a pair of aluminum brackets and mounted with lag screws see (**fig. 2**). It may be necessary to use shims to keep the pulley in a vertical position if the top of the pole is not straight.

Both a support and winch cable are needed. The support cable is a single 70 foot (21.3 m) length of 1/4 inch (6 mm) flexible steel cable. The winch cable is about 75 feet (22.8 m) of 1/8 inch (3 mm) cable. A heavier carriage and antenna would require using a heavier winch cable.

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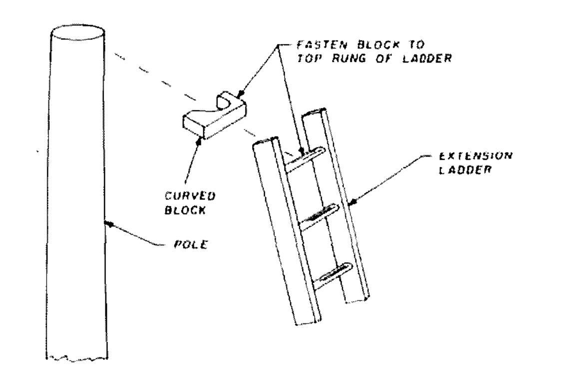


fig. 1. Wooden chock for stabilizing top of ladder.

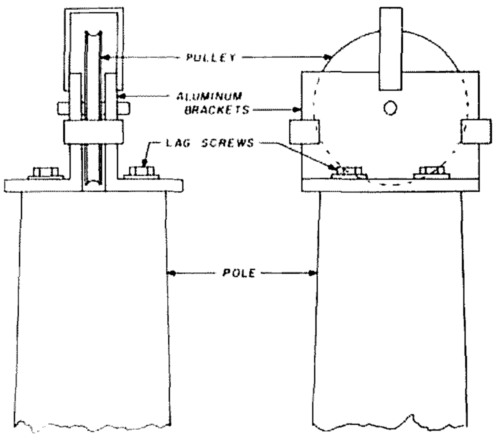


fig. 2. Details of assembly at top of pole.

carriage

The carriage (**fig. 3**) is fabricated from ordinary slotted steel shelving upright strips available at hardware stores. Each strip is 4 feet (1.2 m) long. The steel support cables fit easily in the channels of these strips.

Lay the steel strips on a work bench 10 inches (25.4 cm) apart, the required spacing for the support cables. Bolt two metal supports at right angles to the strips at points one quarter and three quarters of the way up the carriage. Weld the two shelving brackets into the bottom slots of the uprights. Then bolt the bottom shelf to these brackets. Make six clips which will slide onto the upright carriage strips to hold the support cable securely in the groove when the carriage rides up and down the support cables. When mounting the carriage onto the support cables, hold these clips in place with cotter pins.

A spring-loaded plunger (**fig. 4**) is attached on the underside of the bottom plate. When the carriage is in position at the top of the pole, the plunger slides into a mating hole in the pole, to act as a safety catch in case the winch cable breaks. A nylon string is attached to the eye on the plunger assembly so it can be released from the ground.

antenna mast guides

To stabilize the mast, a right angle bracket is installed at the top of the carriage. A hole slightly larger than the diameter of the mast is made in the bracket. A bearing plate is mounted over the hole while the mast is fitted into the rotor. The size of the bearing plate depends on the size of the mast; some measuring and alignment is necessary to assure that the rotor is correctly aligned with the hole in the top bracket and that the mast is straight.

winch

The winch is mounted on the pole at shoulder level. Purchased from Montgomery Ward, mine was strong enough to pull about 1200 pounds (545 kg). It

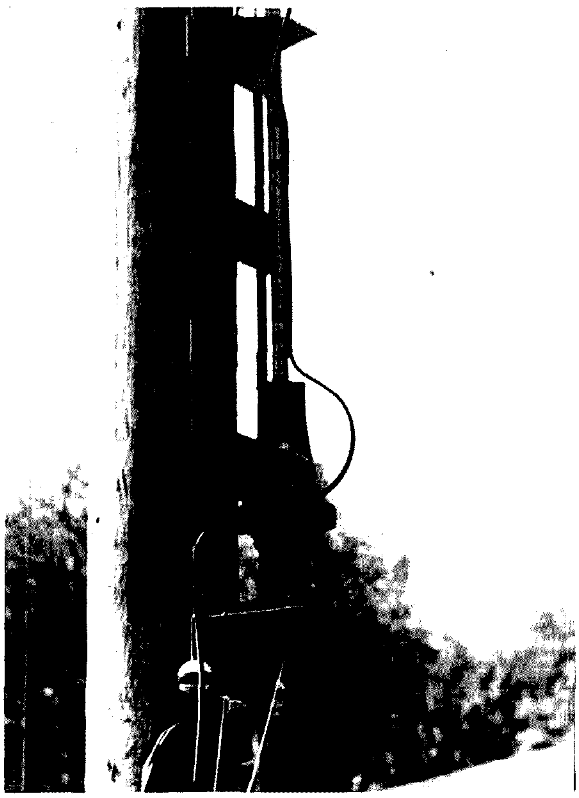


fig. 3. The carriage. Clip can be seen on the left upright. Notice the pipe at the top of the carriage. This was used to wrap the steel cable to the carriage. The two rubber balls mounted at the bottom are shock absorbers.

was spaced far enough away from the pole for the handle to clear and bolted to the pole with two 3/8 inch (9.6 mm) threaded rods. (Each of the two threaded rods should be ground to a point at one end and squared off to accommodate a wrench at the other end. Finished this way, each rod can be screwed into the pole by first drilling a hole slightly smaller in diameter and then using a wrench to turn the rod into the hole. A little grease may make the job easier.) When the rod is in place, cut away the excess length and mount the winch.

vertical guide cable and cable spreader

To install the vertical guide cable on the pole, mount a top support bracket at the top of the pole using two 1/2 inch (13 mm) bolts made from threaded rod (**fig. 5**). The cable spreader (**fig. 5**) is lag bolted to the pole about 6 feet (1.8 m) from the ground. To fasten the turnbuckles install a triangular plate about 3 feet (0.9 m) from the ground using a 1/2 inch (13 mm) rod through the pole (**fig. 5**). Lay the 1/4 inch (6 mm) flexible steel carriage support cable in the top cable bracket and attach the turnbuckles to the ground end of the cables with the cable clamps. After the cable is installed and tightened, the carriage can be mounted and run up and down the pole a few times to assure proper operation.

conclusion

This simplified method of assembly, using inexpensive and readily available materials, can be used to raise antennas to effective working heights.

My antenna stands 20 feet (6 m) high in its lowered position; in the raised position, it stands 40 feet (12 m) high. The pole to which it is attached measures 35 feet (10.6 m).

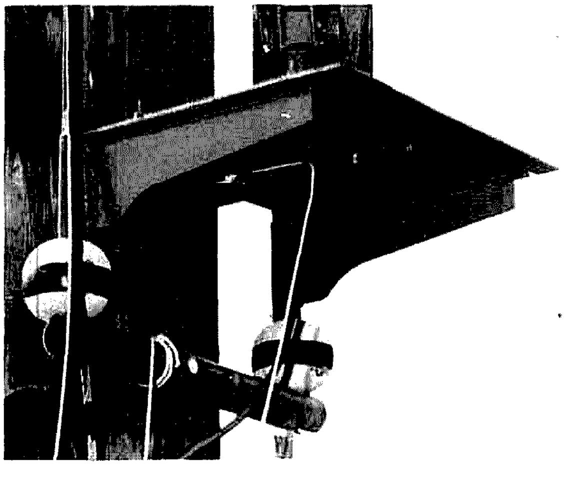


fig. 4. View of plunger assembly.

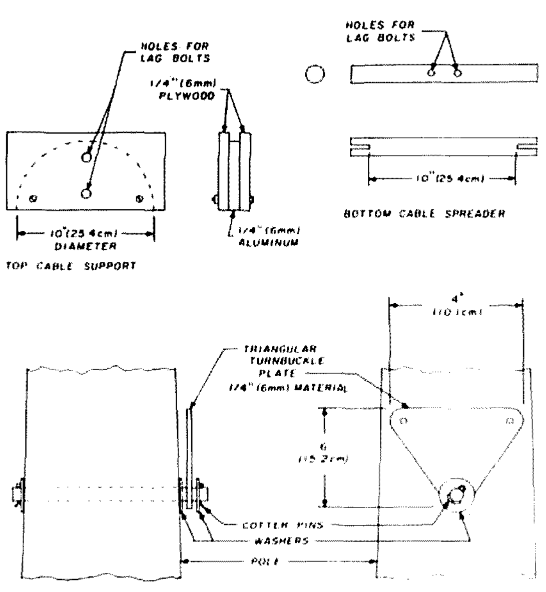


fig. 5. Construction details.

materials list

quantity	*description*
2	*turnbuckles*
4	*cable clamps — 1/4 inch (6 mm) cable*
3	*cable clamps — 1/8 inch (3 mm) cable*
1	*9 inch (22.8 cm) pulley*
1	*hand operated utility winch*
70 feet (21.3 m)	*1/8 inch (3 mm) steel cable*
75 feet (22.8 m)	*1/4 inch (6 mm) flexible steel cable*
30 feet (9.2 m)	*nylon string*
2	*upright steel shelving strips*
2	*shelf brackets to fit strips*

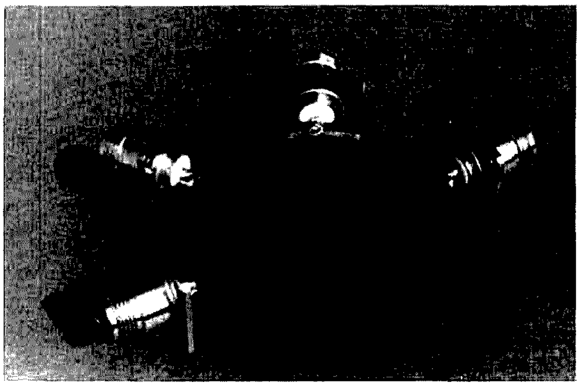
aluminum plate
aluminum mast material
steel plate
lag bolts
threaded rod with nuts (also known as All-thread)
steel metal for brackets, braces, and clamps.

A few details should be noted: be sure to prime and paint any metal parts subject to rust. Lubricate as necessary. Run a ground wire from the support cable and the winch, and install a ground rod at the base of the pole. The transmission line and rotor cable can be run underground to the shack, if you wish. Remember to rotate the antenna only in the raised position.

ham radio

the hybrid ring

Using the "rat race" for power combining, splitting, and coupling



The experimental 1296-MHz hybrid ring with inexpensive type "N" launchers.

The **hybrid ring**, also known as the "rat race," is a device used either to divide or combine power at VHF and UHF. The hybrid ring is easily constructed using double-sided printed circuit boards; tolerances in dielectric thickness and etching accuracy are not critical. The hybrid ring's outputs, in one application, can be either equal amplitude in-phase or 180-degree out-of-phase signals, depending on the input port chosen. Or it can be used as a directional coupler with different power levels available at the output ports.

defining a hybrid

In general, the impedance seen at any port on a hybrid is equal to the characteristic impedance of the transmission line if all of the remaining ports are properly terminated in this same impedance. Each pair of output ports must remain isolated from each other; the input ports must also remain isolated from each other. This is very important when power is to be divided equally to feed two power amplifiers, or when you wish to minimize local oscillator radiation during the combination of two signals (such as in a mixer hybrid.) The hybrid ring, or rat race, is a directional coupler that can be used to sample power trav-

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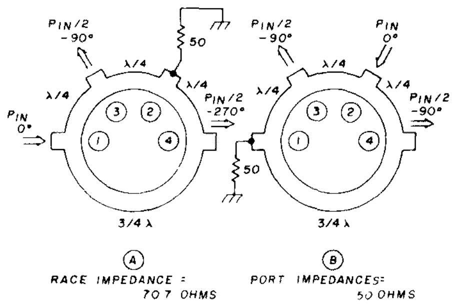


fig. 1. The hybrid ring provides (A) two equal signals 180 degrees out of phase or (B) two in-phase signals.

eling in different directions. It is simple to construct and very tolerant of line-width variations when microstrip transmission line is used. The power ratio at the output ports can be varied by varying the impedances of the interconnecting lines, and the simple hybrid ring can provide a good match and excellent isolation over a ±20 percent bandwidth. The size of the microstrip version of the hybrid limits its use to 432 MHz and above.

The simple hybrid ring consists of a ring of 70.7-ohm transmission line terminated in four places. The top view of the microstrip hybrid ring etched on a printed wire board is shown in **fig. 1A**. Three ports are separated by quarter-wavelength sections. The last transmission line is three-quarters wavelength long, adding up to a total circumference of 1.5 wavelengths.

The hybrid ring is commonly used to split or combine rf power, and if a signal is applied to port 1 of the ring, the power will be equally divided between ports 3 and 4; the phase relationship between the output signals will be 180 degrees. Power incident at port 2, **fig. 1B**, will also be equally divided between ports 3 and 4, but the two output signals will be of similar phase.

Hybrid ring operation can be understood by studying the simple power divider shown in **fig. 2A**. An input signal at port 2 is equally divided at the junction of the two quarter-wave lines. The transmission lines act as impedance transformers whose characteristic impedance is equal to the square root of the product of the end impedances. If the terminations are all 50 ohms, then the quarter-wave line transforms the output load of 50 ohms at port 3 and at port 4 up to 100 ohms at the input, port 2. The parallel combination of the input impedance (100 ohms) to each of the two quarter-wave lines at port 2 is equal to 50 ohms. This divider can also be used as a combiner if two identical signals of equal phase are applied to ports 3 and 4. This power divider is still not a true hybrid, because ports 3 and 4 have only 6 dB of isolation. In other words, a signal applied to port 3 will be 6 dB down when measured at port 4.

Additional transmission lines, **fig. 2B**, transform the simple power splitter into a true hybrid. Any power reflected at output port 3 due to a mismatch arrives at the other output port 4 by two paths. One signal travels one-half wavelength in a clockwise rotation from port 3. The counter-clockwise signal appears at port 4 delayed by a full wavelength. This half-wave difference in arrival time and equal path loss causes the two signals to cancel at port 4, with total cancellation resulting in highest isolation. The reflected signal from any mismatch at port 3 arrives at port 1, in phase from both circular paths, where it is dissipated. This port is designated the isolation port. A detector placed at this port indicates imbalance between the output ports. The input signal from port 2 cancels at port 1 because the clockwise and counterclockwise paths differ by one-half wavelength. If two equal signals with 180-degree phase

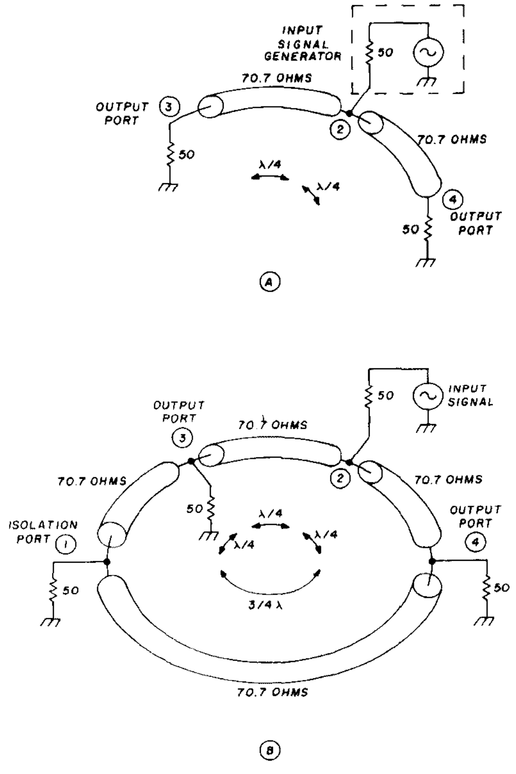


fig. 2. The hybrid ring is formed by (A) an equal split power divider followed by the addition of a cancellation line (B) to provide port-to-port isolation.

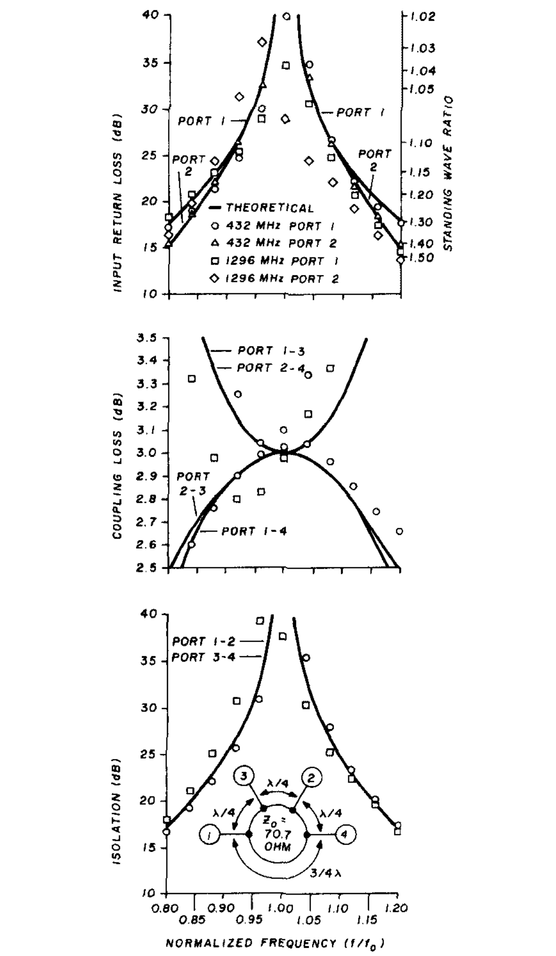


fig. 3. The theoretical response (solid line) of a 70-ohm hybrid ring is compared with experimental results using semi-rigid line.

difference are needed, the input signal can be changed to port 1 and the output taken from ports 3 and 4, as in **fig. 1A**.

70-ohm rat race

The impedance of the ring, or "race," is the port impedance multiplied by the square root of two (*50 ohms* $\cdot \sqrt{2} =$ *70.7 ohms*). The input match of the hybrid is given in terms of return loss, that is, the ratio of reflected power to incident power,

$$\textit{return loss} = -10 \log_{10} (\textit{reflected power}/\textit{forward power}) \quad (1)$$

The theoretical and experimental results of a 1.5-wavelength rat race are shown in **fig. 3**. The experimental results, using semi-rigid coaxial cable to form the race, are shown for 432 and 1296 MHz. The input return loss is greater than 20 dB (SWR $\leq$ 1.2:1) over a 20 percent bandwidth at port 1 or port 2. This means that only 1 percent of the input power is reflected at the input port. The hybrid ring displays an equal power split (3.01 dB) to within 0.25 dB over the same $\pm$ 10 percent bandwidth. This means that at 90 percent and at 110 percent of the center frequency the output power at one port is only 0.25 dB greater (6 percent unbalance) than at the other output port. The isolation between ports 3 and 4 is greater than 20 dB over the same bandwidth. Any mismatch at port 3 causes the reflected signal appearing at port 4 to be at least 20 dB down (1 percent of the reflected signal).

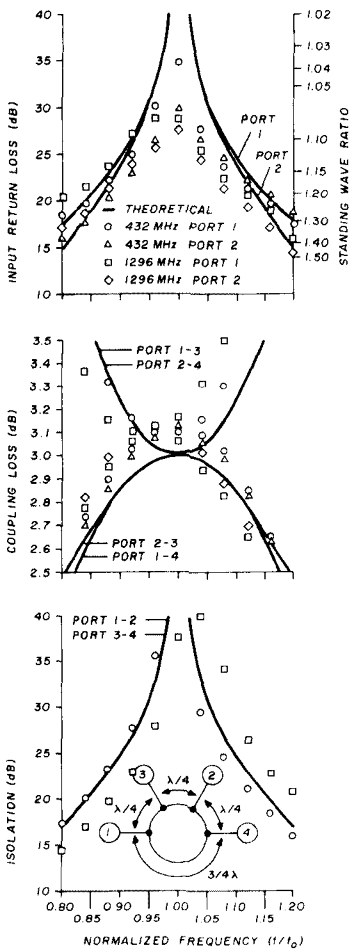


fig. 4. The experimental results using microstrip were also very close to the predicted values at 432 and 1296 MHz.

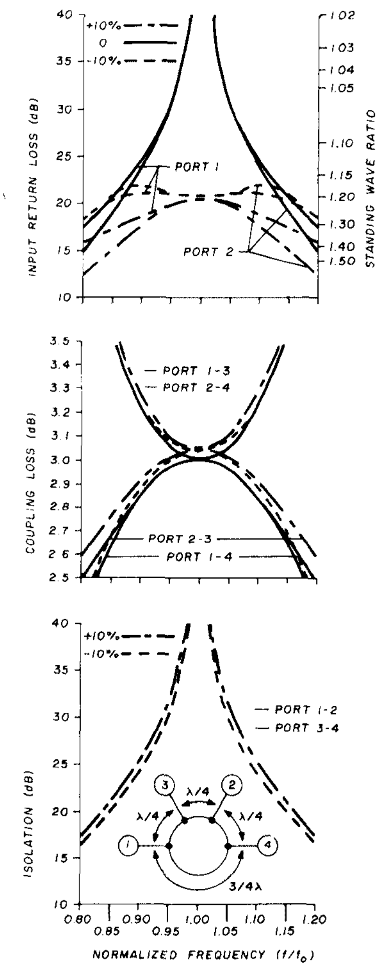


fig. 5. Variations in the impedance of the ring appears to have little effect on the operation of the hybrid ring.

A printed-circuit board version of the hybrid ring was also constructed, with the results shown in **fig. 4**. The width of a microstrip line on one-ounce, 1/16-inch thick, copper-clad Teflon-fiberglass board (e = 2.55) is 95 mils for 70.7-ohm line and 166 mils for the 50-ohm termination leads. The relative velocity of propagation in that board material for a 70.7-ohm microstrip line is 0.700; but it's only 0.688 for a 50-ohm line. The length of a quarter-wave line is thus shortened from its free-space value by this amount. The mean diameter of the ring is simply the average of the inner and outer diameters. The dimensions of hybrid rings for use at several UHF bands are given in **table 1**.

The effects of variations in the impedance of the ring are displayed in **fig. 5**. Variations of ± 10 percent in impedance produce only minor changes; the greatest change was in the input return loss. Still, the hybrid displayed an input return loss greater than 17.5 dB (SWR ≤ 1.3:1) over a 20-percent bandwidth. A variation of 10 percent in ring impedance corresponds to a line width range of from 77 mils to 113 mils about the desired value of 95 mils, for a one-ounce Teflon-fiberglass PC board. This amounts to an almost ± 20 percent variation in the width of the microstrip, much greater than expected.

The PC board version used homemade microstrip-to-coax launchers soldered directly to the ground plane. Type-N female chassis connectors (UG58A/U) were modified by hacksawing a notch, as shown in **fig. 6**. The hacksaw blade was held flat

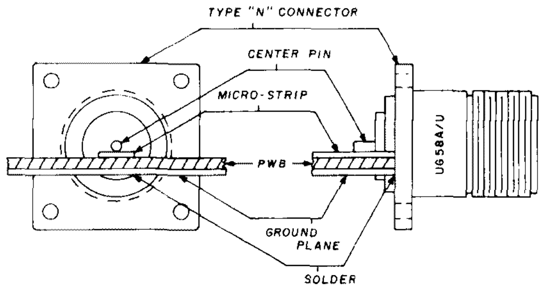


fig. 6. Inexpensive end launchers are formed by modifying common coaxial connectors.

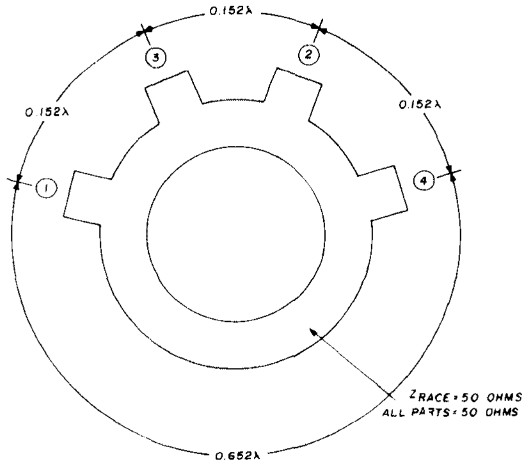


fig. 7. An equal-split hybrid ring can also be formed using 50-ohm transmission line.

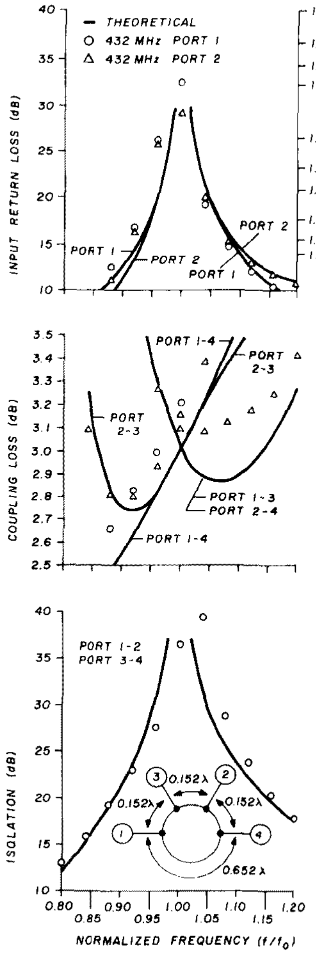


fig. 8. A hybrid ring formed using 50-ohm line microstrip operates over a much smaller bandwidth.

against the center conductor until it penetrated the flange. The PC board was then inserted into the slot and soldered using a 50-watt iron. The measured loss of two launchers mounted to a short microstrip was less than 0.1 dB, the return loss was greater than 32 dB (SWR ≤ 1.05) through 1296 MHz.

50-ohm rat race

If the rat race is constructed of 50-ohm coaxial cable, the cable lengths between each port are shorter, as shown in **fig. 7**.

$$Z\ line = Zo \tan \theta \qquad \textbf{(2)}$$

where Zo is the characteristic impedance of the line (50 ohms) and θ is the electrical length in degrees.

$$70.7\ ohms = 50\ ohms \tan \theta \qquad \textbf{(3)}$$

$$\theta = 54.7\ degrees \qquad \textbf{(4)}$$

The short cable lengths required are 0.152 wavelength; (54.7°/360° = 0.152 wavelength). The cable section between ports 1 and **4** is one-half wavelength longer, 0.652 wavelength. The circumference is only 1.108 wavelengths for the 50-ohm rat race, compared to 1.5 wavelengths for the 70-ohm model. A disadvantage of this lower-impedance hybrid is in the reduced bandwidth, as indicated by the frequency response curves in **fig. 8**.

uneven power-divide rat race

Output power ratios other than 1:1 are possible through selection of different transmission line impedances between the ports. A 10-dB coupler using this approach is illustrated in **fig. 9**. For a signal input P1 at port 1 and in-phase outputs at ports 3 and 4, the value of transmission line impedance is:

$$Z_1 = Z_0 \sqrt{P1/P3} \quad Z_2 = Z_0 \sqrt{P1/P4} \qquad \textbf{(5)}$$

where $P3$ is the output power at port 3 and $P4$ is the power output at port **4**. The sum power from ports 3

table 1. Hybrid ring dimensions for one-ounce, 1/16-inch thick, Teflon-glass board.

frequency (MHz)	70 ohms one-quarter wavelength (inches)	70 ohms three-quarter wavelength (inches)	1.5λ mean diameter (inches)
432	4.78	14.34	9.13
1296	1.59	4.78	3.04
2304	0.896	2.69	1.71

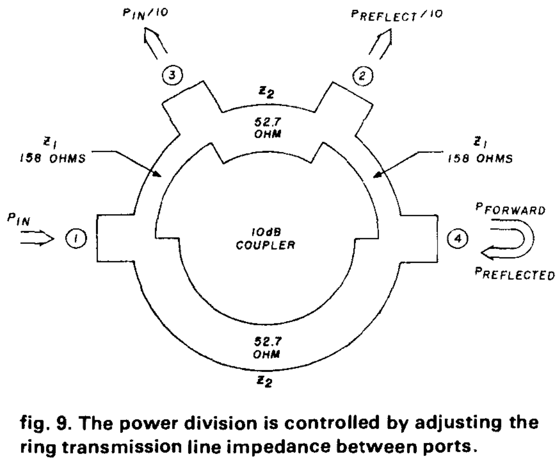


fig. 9. The power division is controlled by adjusting the ring transmission line impedance between ports.

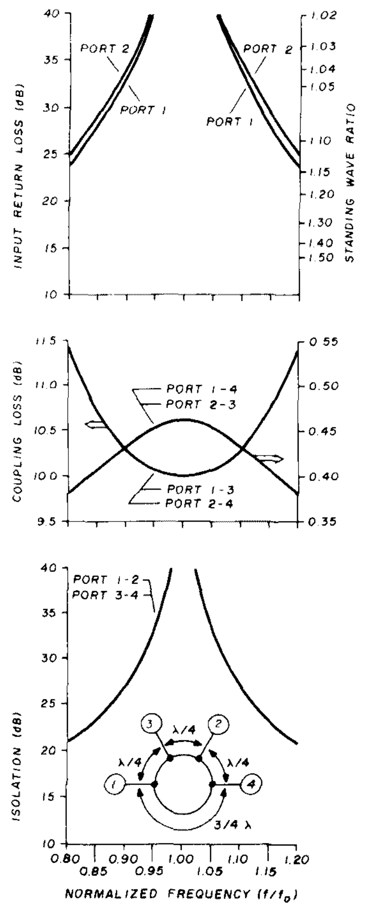


fig. 10. The theoretical response of the 10-dB coupler appears quite good over a 20 percent bandwidth.

and 4 must equal the input power. Constructing a 10-dB coupler, as an example:

$$Z_1 = 50\sqrt{1/0.1} = 158.1\ ohms$$
$$Z_2 = 50\sqrt{1/0.9} = 52.7\ ohms \qquad (6)$$

The output signal at port 3 is a sample of the input signal at port 1 with −10 dB of coupling. To use the hybrid as a directional coupler standing-wave-ratio meter, detectors are placed at ports 2 and 3. Forward power is detected at port 3 and reflected power at port 2. Both signals are sampled 10 dB down from true power. The theoretical response is shown in **fig. 10.** The line width for a 158-ohm line (10 mils) is just too thin, however, for Amateur etching.

applications

The principal use for the hybrid ring is to split or combine power. If more power is needed from a power amplifier than a single transistor can handle, it is necessary to parallel two devices. To maintain stability the two devices must remain isolated from each other. The hybrid performs this function as indicated in **fig. 11.** By using the 180-degree ports, the amplifier operates in a balanced or push-pull manner as seen in **fig. 11A.** The input impedance is effectively four times as great as would be in the case of a single transistor with twice the power-handling capabilities. The case of an in-phase parallel amplifier is shown in **fig. 11B.** Comparison of the insertion loss between the two arrangements (assuming unity gain amplifiers) shows the broader bandwidth response of **fig. 12,** for the push-pull amplifier. If one amplifier should fail, the output power will drop to one-fourth the normal level. The remaining amplifier will deliver one-half its power to the antenna and the remaining one-half to the termination at the isolation port. If the input to either amplifier were to open or short, the input return loss at the hybrid input port 1 would drop to 6 dB (SWR $\leq$ 3:1).

When the rat race is used as a power splitter, each output will have equal amplitude and phase, provided the ports are reasonably terminated. When it is used to combine the output power from two transis-

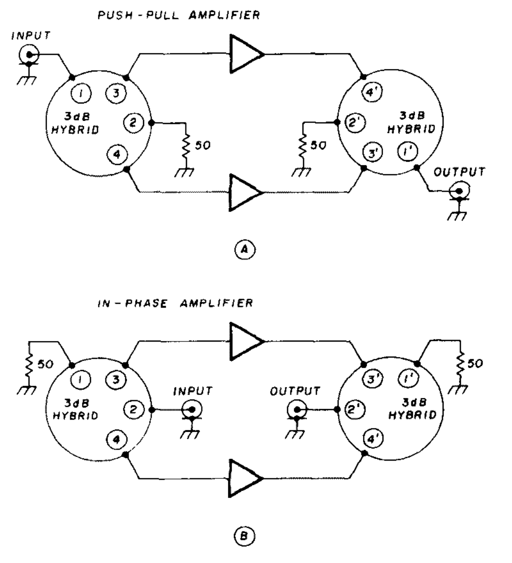


fig. 11. The hybrid ring may be used as a power splitter/combiner for a (A) push-pull amplifier or (B) an in-phase amplifier.

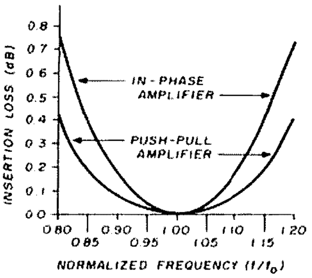


fig. 12. The insertion loss of two hybrids is lower for push-pull connections.

tor amplifiers, the input power delivered to the two input ports may not be equal or in phase. This may be due to differences in transistor gain and internal phase shift. The output power is then less than the sum of the two input powers. The percentage difference (η) of this ideal sum of the two powers is given by:

$$\eta = \left[0.5 + \left(\frac{\sqrt{r} \cos \theta}{r + 1} \right) \right] \times 100 \ percent \quad (7)$$

where r is the power ratio of the two input powers, and θ is the phase angle between them. If the two input signals are in phase but differ in amplitude, the above equation reduces to:

$$\eta = \left[0.5 + \left(\frac{\sqrt{r}}{r + 1} \right) \right] \times 100 \ percent \quad (8)$$

For an input power ratio of 2:1 (3 dB)', the output power will be down only 0.13 dB, or 97 percent of the sum of the two input powers. If the amplitudes are balanced, but the phase of the two input powers differs then,

$$\eta = \left[0.5 + \left(\frac{\cos \theta}{2} \right) \right] \times 100 \ percent \quad (9)$$

For an input phase difference of even 15 degrees, the output power will be down just 0.7 dB, or 98 percent of the available power. For a combination of a power unbalance of 2:1 and a phase unbalance of 15 degrees between inputs, one would suffer a total loss of only 0.2 dB, leaving 96 percent of the original available power.

references

1. Henry S. Keen, W2CTK, "High-Frequency Hybrids and Couplers," *ham radio*, March, 1978, pages 72-75.
2. James R. Fisk, W1HR, "Microstrip Transmission Line," *ham radio*, January, 1978, pages 28-37.

ham radio

a digital audio filter
for CW and RTTY

Build a useful audio notch and bandpass filter

This filter, designed around the National MF10 integrated circuit, consists of CMOS (Complementary Metal Oxide Semiconductor) active filter building blocks. Each block, together with an external clock and a few resistors, can provide different filter functions such as notch or bandpass.

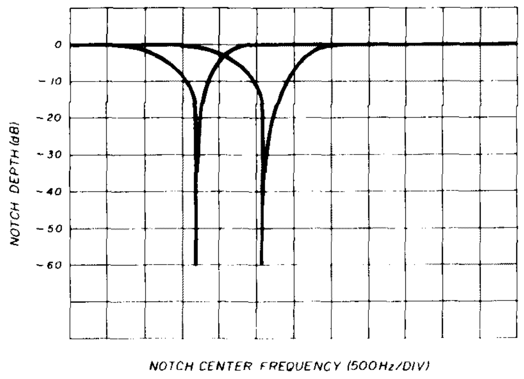


fig. 1. Notch filter response can be translated in frequency by changing the clock rate.

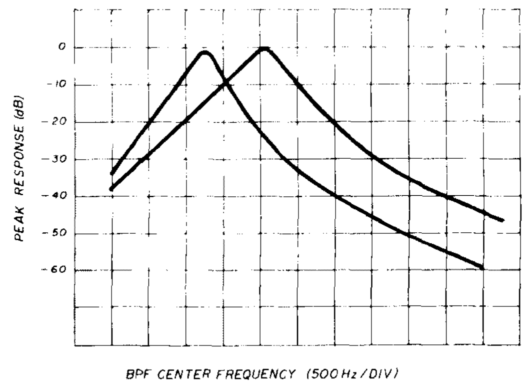
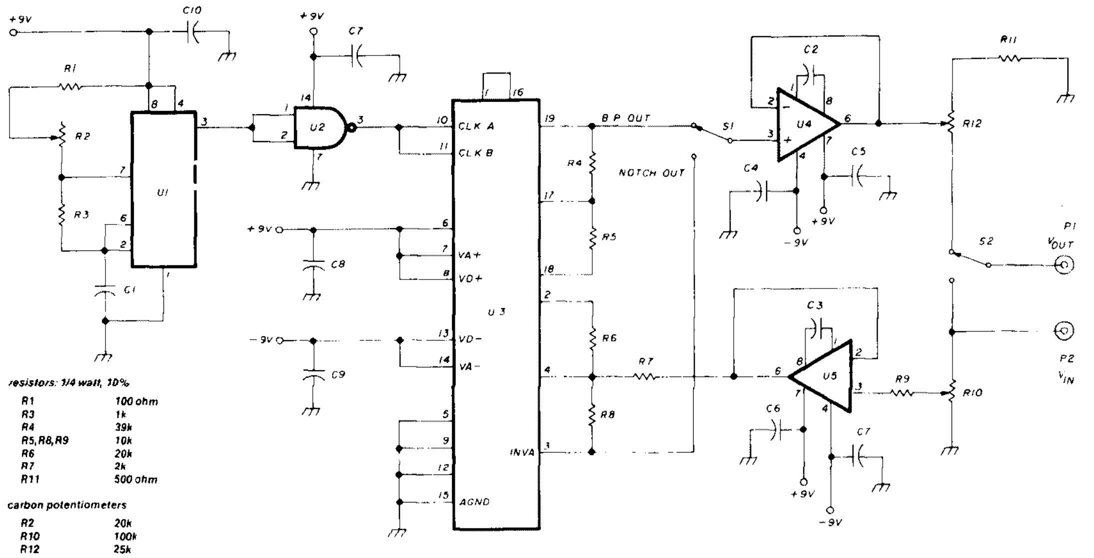


fig. 2. Bandpass filter response at two different center frequency settings.

A major advantage of this type of filter is that the notch and bandpass position is determined by the clock frequency. Therefore, by varying potentiometer R2 the notch or bandpass can be moved in position as shown in the spectrum analyzer photographs in **figs. 1** and **2**. **Fig. 1** indicates a notch filter response (notch depth approximately 60 dB) and **fig. 2**, the bandpass display displaced in two positions. The bandpass filter was adjusted for CW or RTTY operation.

By Don Kadish, W1OER, 135 Barbara Road, Waltham, Massachusetts 02134



resistors: 1/4 watt, 10%

R1	100 ohm
R3	1k
R4	39k
R5,R8,R9	10k
R6	20k
R7	2k
R11	500 ohm

carbon potentiometers

R2	20k
R10	100k
R12	25k

ceramic capacitors

C1	1500 pF
C2,C3	30 pF
C4,C5,C6,C7, C8,C9,C10	0.1 μF

integrated circuits

U1	National 555 timer chip
U2	National CD4011AE
U3	National MF10
U4,U5	National LF365N or equivalent

fig. 3. Audio notch and bandpass filter schematic.

A 555 timer chip, U1, provides the variable clock input. U2 inverts the clock output to providing a TTL level into the clock inputs of U3. A switch, S1, is used to switch from notch to bandpass operation. R12 in conjunction with R11 is a volume control. A speaker or headphones can be directly driven by U4. An input volume control, R10, is used to prevent saturation of the filter stage U3. Once R10 is set it does not have to be adjusted further. U4 and U5 isolate the filter chip from the input and output connections.

constructing and operating the filter

Construction is simple; neither layout nor component values are critical. However, use of decoupling capacitors on all ICs is good practice and minimizes the chance of high-frequency oscillations occurring. All components are mounted on a single-side copper clad vector board.

Filter operation requires connecting V_{IN} to the audio output of a communications receiver through a phono jack and connecting V_{OUT} to a speaker or headphones. Switch S2 to V_{IN} in order to bypass the filter. Adjust the receiver audio gain control for comfortable listening, then switch S2 to V_{OUT} to insert the filter. Adjust the sensitivity control, R10, for comfortable listening volume. Actually the only precaution necessary is to adjust the volume so that clipping of the filter stage does not occur. If clipping does occur, reduce the receiver audio gain until it sounds "clean."

On-the-air tests in the notch filter position gave excellent rejection of adjacent signals. RTTY operation in the bandpass mode is also very simple. Adjust R2 to the extreme end of the potentiometer (the end that accepts the mark and space tones). Except for an occasional adjustment of the volume control, further adjustment is unnecessary.

dc supply voltages needed

Any positive and negative voltage between plus and minus 5 and plus and minus 12 should be satisfactory. Batteries can be used with this device. If low power drain is desired, substitute a CMOS timer (7555) for the 555 timer. All ICs should be of the CMOS type.

ham radio

the ham notebook

carrying case for the IC2AT

To ease the crunch on my pocketbook after buying an ICOM 2AT, I made a leather belt carrying-case. It cost me a total of $11.45, which was a considerable saving, and I had a lot of fun making it!

If you have a leather store in your town, ask them for the scrap leather box. Go through it and find some nice black-dyed leather about one-eighth-inch thick. Draw the outline as shown, then soak the leather overnight to make it soft and workable. After cutting it out, bend the corners by placing it between two blocks of wood and tapping the edge with a rawhide mallet. Don't punch any holes until you have fitted the ICOM inside to see if all the dimensions are right. Trim off the surplus leather on the bent-up edges and punch holes with a No. 2 Rampart punch. Place the leather on the end grain of a block of wood, or it will dull the punch.

The holes should be 3/8-inch apart and staggered on the sides so the lacing will drop down each time you thread through a hole. Use leather lacing with a lace needle. Tie a figure-8 knot at the bottom of the lacing string and pull it up to the first hole. I laced mine starting at the bottom.

quantity	description	approximate price
1	3 feet leather lace	$1.30
1	lace needle	.75
1	No. 2 Rampart punch	2.35
1 pound	scrap leather	4.50
1 dozen	rivets to fit leather	.60

Note: These items can be purchased at Macpherson Leather Corp., 1337 5th Avenue, San Diego, California 92101, if you can't obtain them locally.

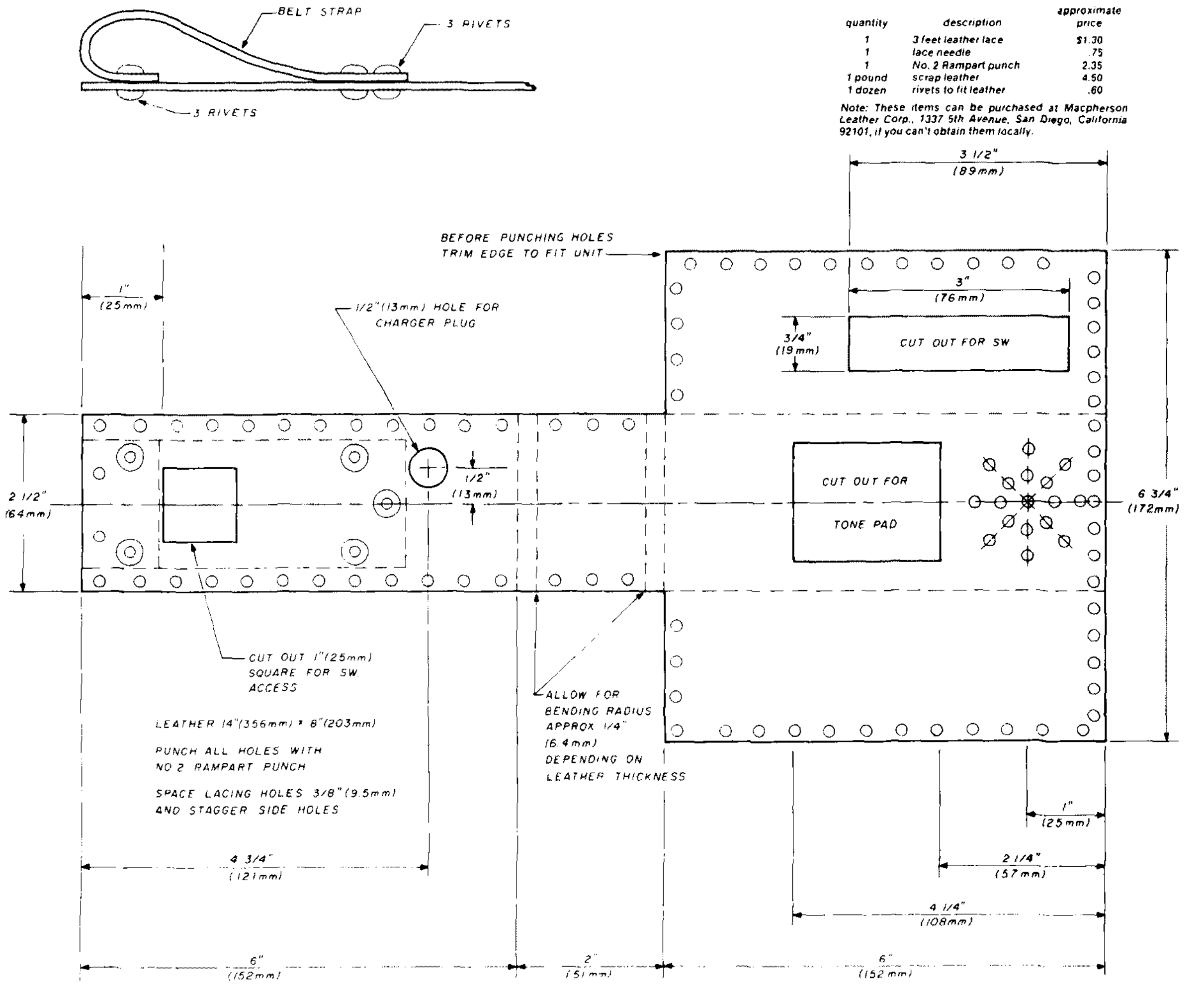


fig. 1. Pattern for belt carrying-case.

The holes for the rivets are also punched with the Rampart punch. (Be sure to buy rivets long enough to go through the leather.) I use a separate belt to carry my ICOM slipped through the belt holder. The belt strap is riveted to the case.

Ed Marriner, W6XM

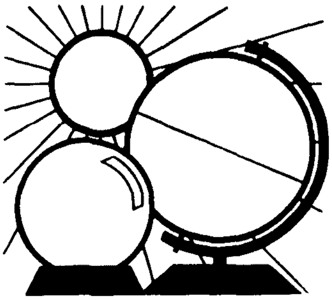
low-duty-cycle tune-up method for transmitters

Having found from sad experience that most final amplifier tubes experience damage and life shortening during tune-up, I decided to use my automatic electronic keyer as a duty-cycle device to cut down tube dissipation during this critical adjustment period. Put your transmitter in the CW position and set your keyer to send dots in the highest speed mode.* Because your transmitter is on only a fraction of the time, your average plate dissipation is low, and you will find it almost impossible to damage your tubes during tune-up. If you work phone often, you can leave your transmitter in the SSB condition and feed the keyer's audio side tone into your microphone input circuit. This is readily accomplished with a simple switch and a small variable potentiometer used to set your input at the desired voltage level. This method will save you the trouble of changing your transmitter mode switches from SSB to CW and back to SSB when you want to tune up. When you use this technique for phone, you avoid the necessity of yelling *AHHHHHHHHHH* into the microphone, a rather unscientific way of establishing a tune-up reference level.

If you do not have an automatic keyer, a relay connected to act as a buzzer with an RC time constant circuit can be used to provide an intermittent on-off low-duty-cycle keying signal. I prefer the automatic keyer, as the tone is a lot cleaner than when using the relay buzzer technique.

William Vissers, K4KI

*Use a dummy load at all times. — Editor



DX FORECASTER

Garth Stonehocker, K0RYW

last-minute forecast

DX conditions for August will depend on which of two solar longitudes is active. Recent information indicates that the lower hf bands (30-160 meters) will be best around the 9th and the higher hf bands (10-30 meters) about the 23rd. However, it's possible that just the reverse will occur. The deciding factor is the 10.7 cm radio flux readings obtained from WWV's broadcast at 18 minutes after the hour. When flux is above 120, use the higher bands. In either case, disturbed periods are expected around the 6th, 15th, and 27th of August with a three day spread on either side of each date.

The moon's perigee will occur on the 8th, with a full moon on the 23rd. The Perseids meteor shower occurs from the 10th to 14th, with its maximum on the 11th and 12th with better than fifty meteors per hour rate. This is an excellent shower.

more on fading

Long duration (slow fade) signal level attenuation was discussed last month. We now consider shorter duration fading or QSB. Two ionospheric conditions (see last month's table, shortwave fadeout and MUF failure) are related to QSB. Listed in this month's table are fading characteristics with possible solutions for each type.

shortwave fadeout (SWF) or ionospheric storm

SWF fading is caused by the geomagnetic field variations that modulate signal levels. The fades are deeper than those caused by solar variations and signal levels take longer to return to normal. The geomagnetic field variations are caused by an in-flux of solar wind particles during the daytime and trapped particles by night. Particles spiral down toward earth following geomagnetic field lines in the polar regions. Particle variations are transmitted to the ions and electrons in the ionosphere and consequently affect the signal level. This signal modulation occurs at its maximum penetration into the ionospheric layer during refraction providing clues of the state of the ionosphere. Higher latitude propagation paths (greater than 60 degrees — Auroral zone) suffer the most attenuation and QSB; transequatorial paths (geomagnetic equator) are next, and mid-latitudes the least during these storms. Night-time QSB is *usually the worst. Auroral QSB is* often fast enough to cause signal "flutter" and is associated with VHF auroral scatter propagation openings.

type of "fade"	speed/ characteristics	best frequency to use	best time to operate
SID	slow	higher band	night/wait 2 hours
PCA	slow/all day	higher band	night
SWF	fast	lower band	day
MUF	slow/deep	lower band	early next day

MUF failure

Consider this: during an afternoon 15 meter opening from the states to South Africa, you note that there's a weak, long-duration fade on signals. After 15 minutes the QSB deepens and becomes even longer in duration. Signal peaks are louder (the result of focusing) and nulls quite a bit weaker. This is explained by the fact that the geomagnetic field separates transmitted energy into three components that travel their separate ways to the receiver. The energy components from the DXer's transmitter are beating against each other, *almost* like zero beating two audio frequencies, until they peak at the MUF, then both decrease in signal strength to a minimum level. This weaker signal is a result of ionospheric forward scatter and has a rough sounding note (see *QST* for January, 1982). Many times this signal is not heard since it is as much as 40 dB weaker than a normally propagated signal (near the MUF).

Another case where the signal takes multiple paths in the ionosphere is at a frequency lower than the MUF. This frequency is low enough (50 percent below the MUF) to propagate by the F_2, F_1, or E layer. A time delay of between 3 to 8 milliseconds occurs between the signals' arrival, causing RTTY pulse elongation errors in addition to its effect on DX signals. It's possible to be too close or too far from the MUF, with its resultant poor propagation conditions.

Often these modes of fading might exist simultaneously. However, when they can be heard separately, useful information is available for predicting near future DX conditions.

WESTERN USA

AUGUST

GMT	PDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	5:00	20	20	30	15	20	10	10	15
0100	6:00	20	20	30	15	20	10	10	15
0200	7:00	20	20	30	15	20	10	10	15
0300	8:00	20	20	30	15	20	10	10	20
0400	9:00	20	20	40*	20*	20	10	10	20
0500	10:00	20	20	40	20	20	10	15	20
0600	11:00	20	20	20	20	20	10	15	20
0700	12:00	—	20	20	20	20	15*	15	20
0800	1:00	—	20	20	20	30	15	15	20
0900	2:00	—	—	20	20	30	15	20	20
1000	3:00	—	—	—	—	30	20	20	20
1100	4:00	—	—	—	—	—	20	20	—
1200	5:00	—	—	—	—	—	—	20	—
1300	6:00	—	—	—	—	—	—	—	—
1400	7:00	—	—	20	20	—	—	—	—
1500	8:00	—	20	20	15	—	—	—	—
1600	9:00	20	20	20	15	—	—	—	—
1700	10:00	20	20	20	15	—	20	20	—
1800	11:00	20	20	20	15	20	20	20	20
1900	12:00	20	20	20	15	20	15	20	20
2000	1:00	20	20	20	15	20	15	15	20
2100	2:00	20	20	20	15*	20*	15*	15	20
2200	3:00	20	20	20	15	15	15*	15	20
2300	4:00	20	20	20	15	15	10	15	20

MID USA

GMT	MDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)	CDT
0000	6:00	20	20	30	15	20	15	15	15	7:00
0100	7:00	20*	20	30	15	20	15*	15	20*	8:00
0200	8:00	20*	20	30	15	30	15*	15	20	9:00
0300	9:00	20	20	30	15	30	15	15	20	10:00
0400	10:00	20	20	40	20	30	15	15	20	11:00
0500	11:00	20	20	40	20	30	20	15	20	12:00
0600	12:00	20	20	20	20	30	20	20	20	1:00
0700	1:00	—	20	20	20	30	20	20	20	2:00
0800	2:00	—	20	20	20	30	20	20	20	3:00
0900	3:00	—	20	20	20	30	20	20	—	4:00
1000	4:00	—	—	20	20	30	20	20	—	5:00
1100	5:00	—	—	—	20	—	20	20	—	6:00
1200	6:00	—	—	—	—	—	—	20	—	7:00
1300	7:00	—	—	—	—	—	—	—	—	8:00
1400	8:00	—	—	20	—	—	—	—	—	9:00
1500	9:00	20	20	20	15	—	—	—	—	10:00
1600	10:00	20	20	20	15	—	20	—	—	11:00
1700	11:00	20	20	20	15	20	20	20	—	12:00
1800	12:00	20	20	20	15	20	20	20	—	1:00
1900	1:00	—	20	20	15	20	20	20	20	2:00
2000	2:00	—	20	20	10	20	20	20	20	3:00
2100	3:00	—	20	20	10	20	15	20*	20	4:00
2200	4:00	—	20	20	15	20	15	15	20	5:00
2300	5:00	—	20	20	15	20	15	15	20	6:00

EASTERN USA

GMT	EDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (Caribbean S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	8:00	20	20	30	15	30	15	15	20
0100	9:00	20	20	30	15	30	15	15	20
0200	10:00	20	20	30	15	30	15	15	20
0300	11:00	20	20	30	15	30	15	15	20
0400	12:00	20	20	30	15	30	15	15	20
0500	1:00	20	20	30	20	30	20*	20	20
0600	2:00	—	20	20	20	30	20	20	20
0700	3:00	—	20	20	20	30	20	20	20
0800	4:00	—	20	20	20	30	20	20	20
0900	5:00	—	20	20	20	—	20	20	—
1000	6:00	—	—	20	20	—	20	—	—
1100	7:00	—	—	—	20	—	—	—	—
1200	8:00	—	—	—	—	—	—	—	—
1300	9:00	—	—	—	15	—	—	—	—
1400	10:00	20	—	15	15	—	—	—	—
1500	11:00	20	20	15	15	—	—	—	—
1600	12:00	20	20	15	15	—	20	—	—
1700	1:00	20	15	15	10	20	20	20	—
1800	2:00	—	15	15	10	20	20	20	—
1900	3:00	—	15	15	10	20	20	20	—
2000	4:00	—	15	15	10	20	15	15	20
2100	5:00	—	15	20	10	20	15	15	20
2200	6:00	20	15	20	15*	20	15	15	20
2300	7:00	20	15	20	10	20	15	15	20

*Look at next **higher** band for possible openings.

band-by-band summary

Ten, fifteen, twenty, and thirty meters will support DX propagation from most areas of the world during daylight and into the evening with a lengthened skip out to 2500 miles (4000 km) per hop. The amount of daylight is still near the yearly maximum, providing many hours of good DXing.

Thirty, forty, eighty, and one-sixty meters are the night DXer's bands. On many nights 30 to 40 meters will be the only usable band because of thunderstorm QRN, but signal strengths via Es short skip may overcome the static, when Es is available. Although Es is available in August, it should be tapering off toward next month. Try the pre-dawn hours for less QRN.

ham radio

new solar index bulletin

A new monthly bulletin from the National Geophysical Data Center provides solar information including:

- daily sunspot number tabulation from previous month
- list of latest predictions of smoothed sunspot numbers
- daily radio flux tabulation of the quiet sun at 9 wavelengths
- non-technical explanatory text

To subscribe, send check or money order ($20 for one-year subscription), payable in U.S. currency and made out to the U.S. Department of Commerce, NOAA/NGDC, to the National Geophysical Data Center, Solar-Terrestrial Physics Division (E/GC2), 325 Broadway, Boulder, Colorado 80303. To charge your subscription on American Express, phone 303-497-6324 or include account number and expiration date with written order.

ham radio

antenna hinge

A method of mounting your beam more easily with less help

For more than thirty years I have used a method of erecting a beam that enables me to put up the antenna either by myself or with a minimum of assistance. I use a simple hinge that permits a Yagi-type antenna to be changed from a horizontal to a vertical position with the removal of a single bolt (**fig. 1**). The hinge is made of two pieces of channel steel or aluminum, with the latter preferred since it's lighter.

The hinge should be as long as the top of the tower is wide. This ensures that when the antenna is tilted to a vertical position the bottom half of the antenna is parallel to the tower. The top half of the hinge should be at least 4 inches (102 mm) wide when you're using it with an antenna boom 2 inches (51 mm) in diameter; a larger-diameter boom requires a wider hinge. This is necessary so that the U bolts, muffler clamps, or the mounting method recommended by the beam manufacturer will give you enough clearance on the bottom of the top piece for properly tightening the nuts.

typical hinge dimensions

bottom half	inches	(mm)	top half	inches	(mm)
length	14	(356)	length	14	(356)
width	5	(127)	width	4	(102)
height	1-7/8	(48)	height	1-1/2	(38)
thickness	3/8	(10)	thickness	3/16	(4.8)

The bottom half of the hinge is positioned with the flat side down (channel up), and the top half, with the flat side up (channel down), is mated with the bottom half. In-line holes are drilled through both pieces approximately 3/4 inch (19 mm) from the ends. These holes should be slightly oversize to freely accept 1/2-inch bolts 6 inches (152 mm) long. This is all that is required at this time.

hinge-to-mast mounting plate

The plate shown in **fig. 2** is cut from 3/16-inch (4.8-mm) steel that measures 17 × 14 inches (43.2 × 35.6 cm). It is almost necessary to have a machine shop fabricate it. In order to reduce weight, 4 × 12-inch (10.2 × 30.5-cm) triangles are cut from each side prior to making a 90-degree bend that provides a horizontal shelf 5 inches wide × 14 inches long (12.7 × 35.6 cm). This leaves a vertical section 14 inches wide at the top, 6 inches wide at the bottom, and 12 inches high (35.6 × 15.2 × 30.5 cm).

The bottom of the hinge can now be mounted to the 5 × 14-inch shelf using four 5/16-inch × 1-inch bolts. The holes are approximately 1-1/2 inches (3.8 cm) in from each end of the shelf and hinge.

parts assembly

Mount a short section of the boom to the top half of the hinge. Mate the two pieces of the hinge and insert the hinge bolts. Remove one of the bolts and the position of the boom can now be changed from horizontal to vertical. Repeat the procedure by replacing the first and removing the second bolt. While this temporary section of boom is in place, two additional holes are required approximately 3 inches (7.6 cm) from each end of the hinge on both sides of the boom. Holes to accommodate 3/8-inch (9.5-mm) bolts are drilled through both pieces of the hinge and mounting shelf, clearing the boom.

After the antenna installation has been completed, the last thing to do before coming down the tower is to install the above bolts (3/8 × 2-1/2-inch) to join the hinge and mounting plate. Without the bolts, wind vibration could damage the hinge. Mark the hinge, top and bottom, so that the ends can always be correctly mated. If reversed, some of the holes might not be in alignment.

Check the antenna for balance before mounting it on the tower. If you balance it well, little effort will be

By J.R. Yost, N4LI, Route 3, Box 342, Mocksville, North Carolina 27028

fig. 1. Beam antenna supported by the antenna hinge.

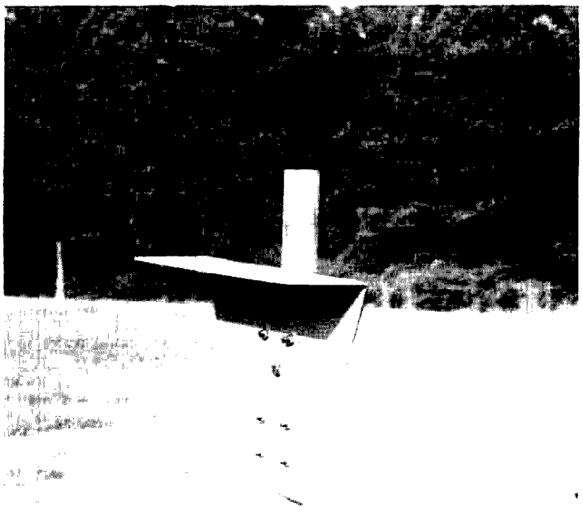


fig. 2. The hinge/mast mounting plate.

needed to change the antenna from a horizontal position to a vertical one.

mounting the antenna

If the antenna weighs more than 50 pounds a gin pole is recommended. The antenna with all elements in place is positioned on the ground at the base of the tower with the boom at a right angle to the tower. The rope from the gin pole is tied to the boom near the end nearest to the tower. By pulling the rope you can stand the antenna on end and lean it against the tower. The rope tied to the boom can now be repositioned to a point 1 or 2 feet above the hinge. A helper on the ground can pull the antenna up the tower, assisted by one man near the top of the tower. The antenna is kept in a vertical position right up to the point where the bottom end of the hinge attached to the boom is at a right angle to the horizontal half of the hinge attached to the mast (**fig. 3**). At this point the holes in the two pieces of the hinge should be

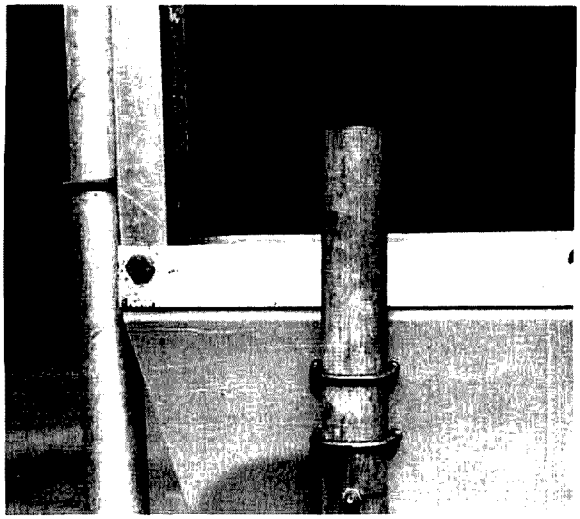


fig. 3. The hinge attached to the mast during erection of the antenna.

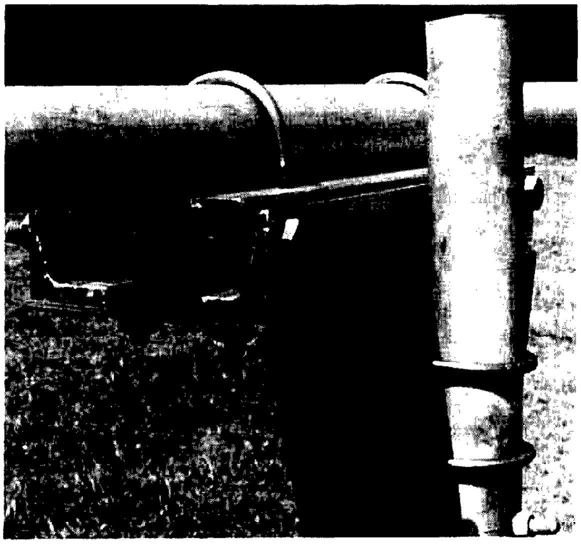


fig. 4. Bolts in place secure the hinge.

aligned and a hinge bolt inserted. With this bolt in place the antenna is secured to the antenna mast (**fig. 4**).

final note

Carefully plan your antenna installation. Write up each step to be taken including the tools needed. Always use a safety belt when working on the tower. Make sure there is no way for the antenna to get near a power line. And always have someone standing by, clear of the tower and the antenna, in case of an emergency.

ham radio

technical forum

Welcome to the ham radio *Technical Forum. The purpose of this feature is to help you, the reader, find answers to your questions, and to give you a chance to answer the questions of your fellow Radio Amateurs. Do you have a question? Send it in!*

Each month our editors will select the best answer received to a question previously posed in Technical Forum. We'll send the writer a book from our Bookstore as a way of saying thanks.

measuring small values

I noticed that K9EBA (*ham radio*, February, 1983) was having trouble trying to measure small values of inductance, so I've decided to add my comments.

I built my own version of the inductance meter described by Ed Marriner, W6XM, in the April, 1982 issue of *ham radio*. Since I wanted to be able to measure values of L over a wide range, I decided the easiest way to do so was to use different rf frequencies, separating that function from the amplifier and meter amplifier. This meant I had to come up with a gain control on the input of the amplifier that did not cause detuning of the tuned circuit in the output of amplifier. (While the circuit shown, **fig. 1**, may not please the purists, it does work. Use no more signal than is necessary, however, in the interest of avoiding harmonic generation in the tuned amplifier.)

I found that sensitivity was poor when trying to measure values of L below 1 μH, so I wound a small air-wound coil in series with the unknown value of L. This increased sensitivity by making for a more favorable L/C ratio at maximum capacity setting of CT and helped tremendously. The size is not critical as long as you can still measure the desired minimum value of L; in my case this was 0.039 μH, as stamped on the case.

As you can see from the enclosed calibration chart for the highest band, 15 MHz (**fig. 2**), it is possible to cover from 3.3 μH down to 0.039 μH with one frequency; lower frequencies can be used to measure larger values of L. 300 kHz can be used to measure as high as 10 mH, the limit for most commonly used inductors, with readily available equipment covering larger values of L.

Do not use the small tuning capacitors normally found in small transistor broadcast receivers at CT. The dielectric tends to wear thinner, throwing calibration off, especially at the maximum capacity setting.

If you use a low cost signal generator for furnishing the frequencies to drive the amplifier, be sure to use minimum setting of output attenuator, since harmonic content is quite high on some of the less expensive generators and can cause false readings when trying to read values of unknown inductors.

I used frequencies as follows: 15 MHz, 3.3 to 0.039 μH; 5 MHz, 33 to 3.3 μH; 1.5 MHz, 500 to 33 μH; 600 kHz, 2.5 mH to 0.5 mH; 300 kc, 10 mH to 1 mH.

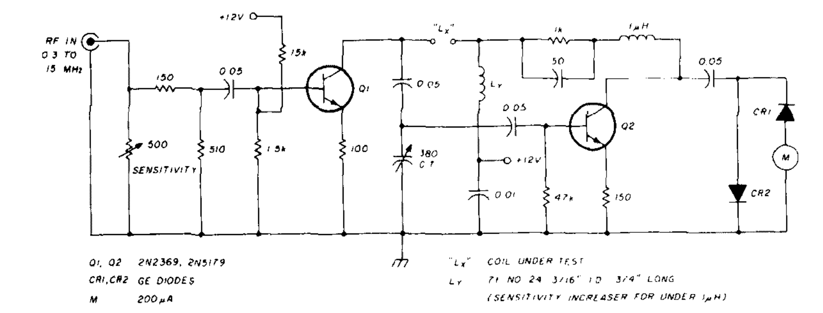


fig. 1. Wide-range inductance meter schematic diagram.

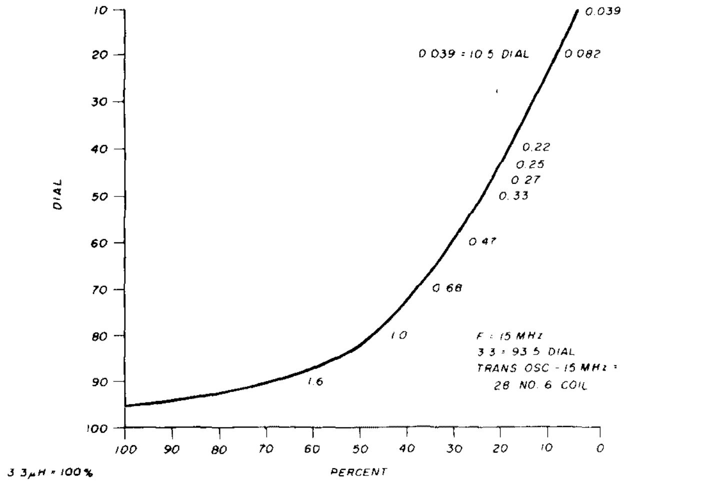


fig. 2. 15 MHz inductance meter dial calibration.

In the schematic and on the calibration chart (**figs. 1** and **2**), to find unknown value of *L*, tune amplifier for maximum on the meter, with 15 MHz signal applied, then pick off the percentage reading that intersects with the dial reading. Multiply the maximum value of *L* (3.3 μH) that equals 100 percent times the percentage thus found; i.e., 1.6 μH equals 49 percent. This is not strictly accurate, since 3.3 does not equal minimum, or zero setting of dial, but it does allow accurate matching, which is usually what is needed. One can tell which *L* is the larger for sure, and that is a help.

On the calibration chart (**fig. 2**) the transfer oscillator 15 megacycle 28 #6 refers to the dipper coil used in the homegrown oscillator I built to use with this thing. Built in a beef stew can, it is an FET Seiler oscillator with buffer which drives the amplifier. (I also used a can to house the inductance measuring device.)

John L. McDonald, W6SDM

✓ 164

✓ 154

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HALLICRAFTERS $$$ Serious collector needs Hallicrafter and other Ham equipment manufactured before 1940 for restoration and eventual museum exhibit. Need Hallicrafters, National, Hammarlund, Patterson, RCA, RME, Grebe, etc. Condition not important. Also need QST mags Vols. I & II and old tubes. All letters answered. Write Dave Medley, WA5YXA, 6621 Duffield Drive, Dallas, Texas 75248.

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WANTED: Schematics-Rider, Sams or other early publications. Scaramella, P.O. Box 1, Woonsocket, RI 02895-0001.

WANTED: Cash paid for used Speed Radar equipment. Write or call: Brian R. Esterman, PO Box 8141, Northfield, Illinois 60093. (312) 251-8901.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

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RADIO WEST! High quality coverage communications receivers with Collins filters and other DX modifications! Catalog 50¢ (refundable). Radio West, 3417 Purer Rd., Dept HR, Escondido, CA 92025 (619) 741-2891.

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GROUND RADIALS WORK. Solve your vertical antenna radial problems with the fantastic ground plane one (GP-1). A 10" diameter, 24-point cast aluminum buss that fits any 2" diameter or smaller mast. Radial problems solved for only $24.95. Send an SASE for photos and brochure. Lance Johnson Engineering, P.O. Box 7363, Kansas City, MO 64116.

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SELL Raytheon 4D32 transmitting tubes new boxed only $30 each. Collection of thirty years receiving tubes $1.00 each. Send list of your needs for availability W5QJT, PO Box 13151, El Paso, TX 79913. (915) 581-2017.

SIGNAL/ONE transceiver, model CX-11A. Solid state 1.5 to 30 MHz, full QSK cw, 150 watt model. Very clean, little use. Contact Mike Russell, at 8668 Celestine Ave., San Diego, CA 92123, or call (619) 278-9333 for details.

SUPER COWW CONTEST programs TRS-80 Model I, III, (IV in III mode) Completely machine language. Logs 3200+ contacts per band. Automatic identification of country and zone from call letters. Dupe Speed 12000+ contacts per second. Screen displays zones still needed, total points, zones, countries, etc. Automatic cw generator with 2 buffers. Log Print program prepares logs and dupe sheets. $39.95 Log Preparation program for hand logs. Similar features to above. $29.95 Free fact sheet and sample printouts. K4SB, 3496 Velma Dr., Powder Springs, Georgia, 30073.

CB TO 10 METER PROFESSIONALS: Your rig or buy ours — AM/FM/SSB/CW. Certified Communications, 4138 So. Ferris, Fremont, Michigan 49412; (616) 924-4561.

CALL FOR PAPERS. 29th Annual VHF Conference, Saturday, October 29, 8 AM to 5 PM with 5:30 dinner, Western Michigan University, Kohrman Hall, Kalamazoo, Michigan. Sponsored by the Electrical Engineering Department. Principle emphasis will be on engineering developments applied to radio communication, design and construction on the frequencies of 30 to 1200 MHz. Authors wishing to present papers should send a synopsis describing the paper to Dr. Cassius Hesselberth, W8FLH, Chairman, Department of Electrical Engineering, Western Michigan University, Kalamazoo, MI 49008. Deadline for submission of synopsis is August 15, 1983. Speakers will be notified of acceptance by August 20.

SIGNAL/ONE transceiver, model CX-7B. One tube in transmitter. Carefully modified power supply, all production updates, receiver RIT, and more. Unit includes spare PA tubes and power supply components. Contact Mike Russell, WB5CSO, at 8668 Celestine Avenue, San Diego, CA 92123, or call (619) 278-9333 for details.

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CUSTOM P/C BOARDS. Send schematic for quotes 1-4 transistors and/or IC's, $15 single board. Orders of 1-10 please enclose payment. Hollan Electronics, P.O. Box 18632, Austin, TX 78760.

Coming Events

ACTIVITIES

"Places to go..."

ALABAMA: The Huntsville Hamfest, Saturday and Sunday, August 20 and 21, Von Braun Civic Center, Huntsville. Free admission. Exhibits, forums, air-conditioned indoor flea market and non-ham activities. Family tours of the Alabama Space & Rocket Center available. Limited camping with hookups available at VBCC. Reserved flea market tables $4/day. Talk in on 3.965 and 34/94. For *information:* Huntsville Hamfest, 2804 S. Memorial Parkway, Huntsville, AL 35801.

ALABAMA: The Central Alabama Amateur Radio Association's 6th annual Hamfest, Saturday and Sunday, August 27 and 28, at picturesque Huntington College Del Champ Student Center, Montgomery. Free admission and parking. Plenty of air-conditioned activities, flea market, DX forum, live RTTY demonstrations and more. Setup 0600, doors open 0800 to 1700 Saturday and till 1500 Sunday. Saturday night dutch treat buffet. Honored guest, G3MLO, Peter Weatherall. Talk in on 146.04/.64. For information or reservations: Hamfest Committee, 2141 Edinburgh Drive, Montgomery, AL 36116 or phone Phil at (205) 272-7980 after 1700 CDST.

DELAWARE: The eighth annual New Delmarva Hamfest, Sunday, August 21, Gloryland Park, 5 miles south of Wilmington. Admission $2.25 advance; $2.75 gate. Tailgating $3.50 with own table. Refreshments available. Talk in on 52 and 13/73. For information and map SASE to Stephen J. Momot, K3HBP, 14 Balsam Rd., Wilmington, DE 19804. Checks payable to Delmarva Hamfest.

ILLINOIS: The Hamfesters Radio Club is having its 49th annual Hamfest and Picnic, Sunday, August 14, Santa Fe Park, 91st and Wolf Road, Willow Springs, southwest of Chicago. Exhibits for OMs and YLs. Famous Swappers

Row. Tickets $3.00 at gate; $2.00 advance. For tickets send check or MO with SASE to Hamfesters, P.O. Box 42792, Chicago, IL 60642.

ILLINOIS: The Chicago Area Computer Hobbyist Exchange and Chicago Amateur Radio Club will hold a joint swapfest, August 21, 10 AM to 4 PM, Triton College, Fifth Avenue, River Grove, IL. For information call 545-3622.

ILLINOIS: The Shawnee Amateur Radio Association's Sarafest '83, Sunday, September 11, John A. Logan College, Highway 13 near Carterville. Displays, flea market, crafts. Doors open 7 AM, free coffee and donuts 7-8 AM. Admission $3.00 at door. Talk in on 146.25/.85; 146.52 simplex or 3.925 SARA Sunday net frequency. For information: William May, KB9QY, 800 Hilldale Avenue, Herrin, IL 62948 or any SARA member.

INDIANA: The Tippecanoe Amateur Radio Association's 12th annual Hamfest, Sunday, Aug 21, Tippecanoe County Fairgrounds, Teal Road and 18 Street, Lafayette. Grounds open 7 AM. Tickets $3.00. Large flea market, dealers, fun, refreshments. Talk in on 13/73 or 52. For information/tickets: Lafayette Hamfest, Route 1, Box 63, West Point, IN 47992.

KENTUCKY: The Bluegrass Amateur Radio Society will sponsor the Central Kentucky ARRL Hamfest, Sunday, 8 AM to 5 PM, August 14, Scott County High School, Longlick Road and US 25, Georgetown. Tech forums, awards, exhibits. Free outdoor flea market space. Tickets $3.50 advance, $4.00 at gate. For information/tickets: Edward B. Bono, WA4ONE, P.O. Box 4411, Lexington, KY 40504.

MICHIGAN: 7th annual Five County Swap-N-Shop, Sunday, August 28, Bentley High School, 1150 Belsay Road, Flint. 8 AM to 3 PM. Sponsored by the Genesee County RC, Bay Area ARC, Lapeer County ARRC, Saginaw Valley ARA, Shiawassee ARA. Refreshments, computer forum, trunk sales, free parking. Tickets $2.00 advance; $3.00 at door. Children under 12 free. For table reservations: Bill Cromwell, KU8H, 1204 Overland Drive, Lennon, MI 48449. (517) 288-5046.

MISSOURI: The St. Charles Amateur Radio Club's 8th annual Hamfest, August 28, Wentzville Community Club. Large open air flea market; indoor air-conditioned exhibits. Contests, food and fun. Parking $1 per car. Admission: $1 each, 4/$3 advance. At door $1.50 each, 4/$5. For tickets/information: SCARC Hamfest '83, PO Box 1429, St. Charles, MO 63301.

MISSOURI: The 2nd annual Ozarks ARC Congress and Swapfest, Sunday, September 11, Monett City Park, U. S Hwy 60 and MO Hwy 37, Monett. Free admission. No charge for swappers or tailgaters. Covered dish picnic. Talk in on 146.37/.97, 146.52 and 7.250. For information: OARS, Box 327, Aurora, MO 65605.

NEW JERSEY: The fifth annual Gloucester County ARC Ham/Comp Fest, Sunday, August 28, Gloucester County College, Tanyard Road, Sewell, 8 AM to 3 PM. Setup 7 AM. Seminars, contests FCC exams. Tickets $2.00 advance; $2.50 door. Tailgaters and dealers $3.00 per parking space. Talk in on 146.52, 147.78/18 and 223.96/224.36. For information: GCARC Hamfest Committee, PO Box 370, Pitman, NJ 08071. (609) 456-0500 or 330-4041 days. (609) 629-2064 evenings.

NEW JERSEY: The Ramapo Mountain ARC's 7th annual flea market, August 20, Oakland American Legion Hall, 65 Oak Street, Oakland, 20 miles from GW bridge. Admission $1.00, non-ham family members free. Indoor tables $6.50; tailgating $3.00. For information: Tom Risseeuw, N2AAZ, 63 Page Dr., Oakland, NJ 07436. (201) 337-8389 after 6 PM.

NEW YORK: The annual Finger Lakes Hamfest, August 27, Trumansburg Fairgrounds, Rt 96, 12 miles NW of Ithaca. 8 AM to 5 PM. Admission $2.00 at gate. Flea market commercial exhibitors, boat anchor auction, refreshments and craft show. Talk in on 37/97 and 52. For information: Dave, W2CFP, 866 Ridge Road, Lansing, NY 14882.

NEVADA: Pacific Division ARRL Convention hosted by the Wide Area Data Group, August 19-21, at the MGM Grand Hotel, Reno. Tickets (for convention only) $7.50 advance; $10.00 at the door. With banquet and after-dinner sessions $35.00 advance; $37.50 at the door. Roy Neal K6DUE is guest speaker at the banquet along with Dave Sumner, ARRL General Manager and Vic Clark, ARRL President. Besides convention forums, swapmeet and exhibits, there'll be plenty of opportunity for sightseeing in this beautiful area. Take a cruise on Lake Tahoe, visit Harrah's Automobile Museum and, of course, there's Reno itself — "The Biggest Little City in the World".

OHIO: The 41st Findlay Radio Club's Hamfest, Sunday, September 11, Hancock Recreational Center, 3430 North Main St., Findlay. 6:30 AM to 5 PM. Admission $3.00 advance; $4.00 door. Arena tables $6.00 each. Outdoor flea market space $6.00. Talk in on 147.150/.75. For information: Dave Fleming, N8EOZ, PO Box 519, Findlay, OH 45840.

PENNSYLVANIA: The 46th annual South Hills Brasspounders and Modulators Hamfest, August 7 from 9 AM to 4 PM, South Campus of the Community College of Allegheny County, Pittsburgh. Tickets $3.00 each or 2/$5. Computers, OSCAR, ATV demos. Flea market. Talk in on 146.13/73 and 146.52 simplex. For information: Andrew Pato, 1433 Schauffler Drive, W. Homestead, PA 15120.

PENNSYLVANIA: The Central Pennsylvania Repeater Association's 10th annual Hamfest/Computerfest, August 28, adjacent to Hersheypark, Chocolate Town, USA. Registration $3.00. Wives and children free. Special reduced admission to Hersheypark for registrants' families. Indoor dealer and flea market, outdoor tailgating. 10 ft. indoor space $8.00 each; 8 ft. tables $4.00 each; single electric plug $1.00 each. Talk in on 145.47, 146.76 and 146.52. For information/registration: Timothy R. Fanus, WB3DNA, Hamfest Reservations, 6140 Chambers Hill Road, Harrisburg, PA 17111. (717) 564-0897 noon to 8 PM.

PENNSLYVANIA: The Tioga County ARC's 7th annual Hamfest, Saturday, August 20, 0800 to 1600, Island Park, Blossburg. Flea market, food and more. Talk in on 146.19/.79 and 146.52. For information: Tioga County ARC, PO Box 56, Mansfield, PA 16933 or John T. Winkler, WB3GPY, RD #2, Box 269, Wellsboro, PA 16901 on .19/.79.

PENSYLVANIA: The Mid-Atlantic Amateur Radio Club's annual Hamfest, Sunday, August 14, 9 AM to 4 PM rain or shine, Route 309 Drive-In Theater, Montgomeryville. Admission: $2.50 plus $1.00 for each tailgate space. Tailgate setup 8 AM. Plenty of parking, refreshments and more. Talk in on WB3JOE/R, 147.66/06 or 146.52 simplex. For information: write the Club, PO Box 352, Villanova, PA 19085.

TENNESSEE: The Lebanon Hamfest sponsored by the Short Mountain Repeater Club, Sunday, August 28, Cedars of Lebanon State Park, US 231, Lebanon. Outdoors only. Exhibitors bring own tables. Refreshments available. Talk in on 146.31/146.91. For information: Morris Duke, W4WXQ, 210 Disspayne Drive, Donelson, TN 37214.

VIRGINIA/WEST VIRGINIA: The Bluefield Hamfest sponsored by the East River Amateur Radio Club, Sunday, August 28, Brush Fork Armory-Civic Center, 9 AM to 3 PM. Admission $3.00. Large indoor flea market, dealers, computers, satellite TV and more. Paved parking, refreshments. Talk in on 144.89/145.49 and 146.52 simplex. For information: Don Williams, WA4K, 412 Ridgeway Drive, Bluefield, VA 24605.

VERMONT: BARC International Hamfest, August 13 and 14.

WASHINGTON: The annual Hamfair of the Radio Club of Tacoma, Saturday and Sunday, August 13 and 14, Olson Auditorium, Pacific Lutheran University. Seminars, flea market, exhibits, contests, dinner and loggers breakfast. Tickets $5. Contact Grace Teitzel, AD7S, 701 So. 120th, Tacoma, WA 98444. (206) 564-8347.

RADIO EXPO: Sponsored by the Chicago FM Club, Saturday and Sunday, September 24 and 25, Lake County Fairgrounds, Routes 120 and 45, Grayslake, Illinois. Flea market opens 6 AM. Exhibits open 9 AM. Indoor flea market tables available at $5.00 per day. Tickets $3.00 advance, $4.00 at gate, good for both days. Seminars, tech talks, ladies' programs. Talk in on 146.16/76, 146.52 and 222.5/224.10. For information: SASE to Radio Expo 83, Box 1532, Evanston, IL 60204 or (312) 582-6923.

ALABAMA: Hospitality Hamfest, sponsored by the Mobile Amateur Radio Club, September 10 and 11, Al's Party Palace, 2671 Dauphin Island Parkway (1 mile off I-10). Doors open 9 AM. Admission free. Exhibits, swap tables, YL activities. Talk in on 146.22/82. For information: Jim Wilder, N4GUC, 424 Cody Road South, Mobile, AL 36609. (205) 343-7365.

OPERATING EVENTS

"Things to do..."

AUGUST 10-13: The Cuyahoga Falls Amateur Radio Club will operate W8VPV/8 from the All-American Soap Box Derby, Akron, Ohio 2300 to 0200Z, August 10-12 and 1500 to 2200Z, August 13. Frequencies: 3945, 7265, 14.240, 21.365 and 28.595 ± QRM. Novices and Technicians look 10 kHz up from lower band edge on the hour. *For certificate send 2 units of postage to:* Cuyahoga Falls ARC, PO Box 6, Cuyahoga Falls, OH 44222.

AUGUST 13: The Marin Amateur Radio Club will operate W6SG to commemorate the 50th anniversary of its founding. Operation will begin 1700 UTC and continue for the day. Bands, modes and frequencies: 20M CW, 14.065; 20M phone, 14.265; 15M CW, 21.065; 15M phone 21.365; *40M CW, 7.065; 40M phone 7.265.*

AUGUST 13: The Tri-County Wireless Group will operate N8COY from the famous Grand Hotel's "Longest Porch in the World" on Mackinac Island, Michigan. Operation 1500-2300Z. SSB 7.280, 14.280, 21.380, 28.580 and FM 147.480. For a special QSL send QSL and regular SASE to N8COY.

AUGUST 13 AND 14: The Bergen Amateur Radio Association will operate K2TM from 1500 to 2400Z to mark BARA's 20th anniversary. Frequencies: 7.235, 14.275, 21.380, 28.610, 146.520. For a certificate send large SASE and QSL to K2UFM, Warren P. Hager, 31 Forest Drive, Hillsdale, NJ 07642.

AUGUST 20 AND 21: New Mexico QSO Party sponsored by the Albuquerque DX Association, 1800Z Saturday to 2100Z Sunday. Suggested frequencies: Phone 1.835, 3.985, 7.230, 14.280, 21.370, 28.570, 147.510. CW 1.805, 60 kHz from low end. Novice 25 kHz from low end. Work stations once per band/mode. Repeat QSO's allowed if NM station changes counties. Exchange: NM stations send signal report and county. Others send signal report and state, Canadian province or DXCC country. Scoring: Two points for each phone QSO, three for each CW QSO. NM times total of NM counties, states, provinces, countries worked. Others times total of NM counties worked. Single and multi-operator. Awards: Plaque to highest scoring NM station and non-NM station in each category. Mail entries by October 1 (include large SASE) to N5HH, *Ed Graham, 12449, Regent NE, Albuquerque, NM 87112.*

AUGUST 27 AND 28: The Rochester Amateur Radio Association (RARA) will operate an Amateur Radio station at Camp Good Days and Special Times, Camp Onanda, Canandaigua Lake, New York. This is a special camp for children who have cancer. Frequencies and bands: 80 meters, phone, 3900 and 3925 MHz, CW 3525 to 3550 MHz. *40 meters, phone, 7230 and 7250; CW 7025 and* 7050 MHz. 15 meters, phone 21350 and 21375 MHz; CW 21025 and 21075. Also 2 meters FM locally. RARA members will use club call sign K2JD. SASE for a special certificate.

SEPTEMBER 10: The West Alabama Amateur Radio Society (WAARS) will operate special event station W4WYP *in commemoration of the birthdate of college* football's winningest coach, Paul "Bear" Bryant. 1300 to 2400Z. Frequencies: bottom 25 KC on General 40-15 meter phone band. Novices bottom 25 KC of Novice band. For a handsome commemorative certificate send $1 and large SASE to West Alabama ARS, PO Box 1741, Tuscaloosa, AL 35403.

SEPTEMBER 10 AND 11: Cray Valley Radio Society's 13th SWL Contest, 1800 GMT to 1800 GMT. Up to 18 hours logging allowed. Rest period must be shown. Multi-operators may log continuously. Contest open to anyone. Two sections and two categories. Phone and CW. Single/multi-operator. Bands: 1.8, 3.5, 7, 14, 21 and 28 MHz. Scoring: one point for each station heard times number of different countries heard on each band. A list of countries must be furnished and a separate log for each band. Call areas of the U. S. A., Canada and Australia will each count as a separate country. No CQ or QRZ calls allowed. Log sheets are available for large SASE from Owen Cross, G4DFI, 28 Garden Ave, Bexleyheath, Kent DA7 4LF. Entries to contest manager, G4DFI at above address NLT October 31, 1983. Certificates award*ed at the discretion of the Cray Valley RS.*

SEPTEMBER 10 AND 11: The Starved Rock Radio Club will operate club station W9MKS from their clubhouse in Oglesby, LaSalle County, Illinois. Operation will be on all Amateur bands. A special QSL is being designed for this occasion in celebration of 50 years of Amateur Radio in Central Illinois.

SEPTEMBER 13-17: The Southern Counties Amateur Radio Association (S.C.A.R.A.) is planning to have a special events station during the Miss America Pageant. Check September Ham Radio for details.

product

REVIEW

Alden's weather chart recorder kit

The weather — it's the one thing that affects us all equally. On a tranquil summer afternoon you can watch the sun heat the atmosphere and turn cottony white cumulus clouds into brutal cumulonimbus thunderstorms. Just a few months later comes the chill of winter. Whatever the conditions, there's little you can do except be aware and try to prepare for what comes next.

If you're like me, the evening weather report is probably the most important part of the day's news. But however encouraging or discouraging the prognosis is, there's always a lingering doubt about its accuracy.

There *are* other ways to forecast the weather in your immediate vicinity than watching the evening news or checking the color of the sunset.

Alden Electronics, one of the largest suppliers of radiofacsimile recorders, contacted us recently and asked if we would be interested in building and using one of their new Model 9321 Weather Chart Recorder kits. Because I've been interested in weather science since high school, I said I'd be happy to.

Alden sent a number of brochures describing their commercial product line. From what I saw and read, their capabilities seem to cover just about every aspect of the weather chart recorder field, from high frequency and telephone to direct-from-satellite reception.

A brief explanation may help clarify how and why radiofacsimile charts are important. Ships at sea can't tune in Willard Scott from the Today Show for the morning forecast. They depend solely upon broadcasts from fifty government-sponsored stations in more than twenty-five countries for status charts, forecast charts, satellite pictures, and ice flow charts, to name just a few sources of weather information. (Other sources include surface weather charts, surface wind analysis and prognosis, wave analysis, oceanographic charts, sea ice charts, and significant weather depiction charts.) Sometimes life-or-death decisions are based upon the information provided from these sources; for example, with up-to-date weather information, a ship at sea can steer safely away from a hurricane's path.

the basics

The Alden 9321 is designed using commercially tested and proven circuitry. Relatively speaking it is mechanically very simple and has all the latest in state-of-the-art electronics. It is packaged in a rugged plastic case designed to withstand the rigors of a hostile environment. And, unlike other facsimile equipment I've used, the Alden unit is relatively quiet and free from smoke odor and fumes. (I remember reading about a facsimile recorder that required an extensive venting system to remove the foul odors that accompanied its operation.) Light and compact, it requires only a stable, general-coverage SSB receiver to be put on line.

The 9321 uses two motors to provide stylus scanning and paper feed. The first motor is a tachometer-controlled dc motor that sweeps the stylus across the paper while the second motor slowly plays the paper out of the easy-to-change paper cassette. The stylus scans across the paper; when an image is transmitted, a slight electrical current flows through the paper and causes the image to appear. The reason the Alden unit does not produce unpleasant odors is that its specially-designed moistened paper does not burn, but instead uses electron deposition to produce the image.

the kit

As I unpacked the kit, I discovered that the shipping box is divided into twelve clearly marked compartments numbered to correspond to the twelve-step, easy-to-follow, illustrated instruction manual. Parts for each subassembly are clearly inventoried and identified to make the construction process as simple as possible. To ensure that the parts are not damaged in transit, they are carefully stored in plastic bags and wrapped in protective foam.

Assembly is basically mechanical, not electronic. Each step is carefully detailed, diagrammed, and explained in the forty-two page instruction manual. All of the electronics for the Model 9321 kit are factory assembled and tested, thereby taking advantage of the similarity to Alden's line of commercial facsimile recorders. Assembly involves nothing more than selection of parts, installation, and in some cases, physical alignment. All interconnections are either through ribbon connectors or sturdy nylon shell pin connectors. The only electrical work is the installation of the LEDs and switches in the control panel assembly.

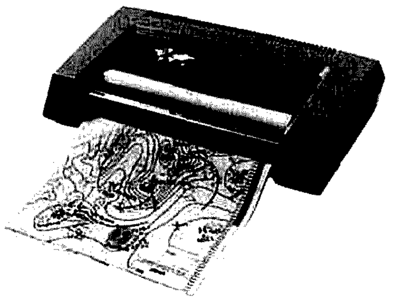
Some kit builders may be disappointed that the electronics are preassembled and tested. But my feeling is that kit builders benefit from Alden's experience . . . and for most of us, anyway, the real reward comes when the unit is turned on and the first charts start to come off the recorder.

It took ten hours to put this kit together — not in one sitting, but rather during evenings within a one week span. Before turning the recorder on I ran a few tests to ensure that assembly had been done according to the manual. (Testing procedures are fully described with step-by-step instructions to make the process as simple as possible.)

Alden went to the trouble to include two very important and helpful booklets as part of their instruction manual. *The Worldwide Marine Radio Facsimile Broadcast Schedule* is a complete compilation of stations from around the world, and includes frequencies, transmission times, and schedules. Here on the east coast, the strongest U.S. station is the Navel Eastern Oceanography Center (NAM) in Norfolk, Virginia. NAM transmits weather data that covers the area east of the Mississippi River including the Gulf of Mexico, the Caribbean, and the northern half of the Atlantic Ocean. In addition to various status maps, they also transmit detailed computer-enhanced photos taken from the GOES geostationary satellites. These photos are taken using either infrared or visible light so they can be of use day or night.

Alden has also included a reprint of the Naval Eastern Oceanography Center's *Facsimile Product Guide*. The facsimile service from Norfolk can be broken down into four separate categories: atmospheric analysis, atmospheric prognosis, oceanographic analysis, and oceanographic prognosis. Each service is fully explained with a number of illustrations and examples included to assist in interpretation of information received.

As I mentioned earlier, there are over fifty stations transmitting up-to-date weather information from twenty-five different countries. Most of these stations transmit at 120 scans per minute at an Index of Cooperation of 576.



(The Index of Cooperation is an internationally agreed upon standard for expressing compatibility between transmitting and receiving equipment.) A few charts are transmitted at 288 IOC, but most are transmitted at 576; consequently, you'll be able to get good quality pictures from the Alden 9321 for the majority of facsimile frequencies. (The 288 IOC charts will be compressed when received on a 576 IOC recorder, but are still usable.)

tuning

Facsimile broadcasts are normally transmitted on upper sideband. After you connect the

unit to 120 volts ac and the audio output from the receiver, tune the receiver approximately 1900 Hz lower than the facsimile station's frequency to correctly position the facsimile sensing circuitry. For example: to correctly tune NAM on 3357, tune the receiver down to 3355.1 kHz. When you are correctly tuned, the two LEDs will be flashing. The green LED corresponds to white and is usually the one that will be lighted the most. The red LED lights only when a black image is transmitted.

At any given moment of the day, somebody, somewhere, is transmitting weather information. So it's likely that whenever you turn your unit on, you'll be receiving a chart — but perhaps not properly framed. If you turn your unit on and find that you've missed the framing signal, all you have to do is push the framing button and keep it depressed long enough to center the chart correctly.

The unit also incorporates an auto-start and auto-stop feature so there's no need to be present during the transmission of charts. Each broadcasting station transmits a signal shifting between 1500 and 2300 Hz at a 300 Hz rate for three to five seconds before beginning transmission of a chart. This tone triggers the auto-start. The framing signal at the beginning of each chart is a 1500 Hz tone for approximately 40 scan lines interrupted once each scan line by a 2300 Hz pulse. The auto-stop signal at the end of each chart is a signal shifting between 1500 and 2300 Hz at a 450 Hz rate for three to five seconds.

The first two charts I received were an atmospheric analysis chart and a GOES satellite picture. These gave clear details of an approaching storm, including its precipitation and cloud cover. They confirmed the weekend forecast, which was bad. I was fascinated by the upper air or steering current charts which came later on.

As I mentioned before, it's fun to try to second-guess the local weatherman. As it turned out, he was right. But I'm looking forward to being able to watch the weather as it develops and make my own forecasts.

Facsimile hasn't yet caught on as a popular mode of Amateur communication. Facsimile devotees are much like the RTTY and SSTV gang — they're relatively few in number, but very interested and quite active. The new Alden 9321 will go a long way toward popularizing facsimile in the Amateur ranks. I'm sure it will be only a matter of time before enterprising and knowledgeable tinkerers will be hard at work modifying this equipment to make it do more than even Alden could have imagined.

I've done a lot in Amateur Radio over the past sixteen years from 160 meter DXing to 2-meter fm. I can truthfully say that I've really enjoyed using the Alden 9321 Recorder. I'm sure you will, too.

For more information, contact Alden Electronics, Washington Street, Westborough, Massachusetts 01581. RS#301

N1ACH

NEW products

connector connection

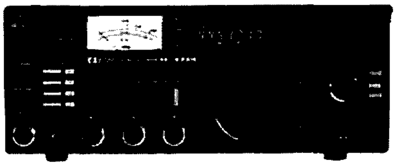
Nemal Electronics, Miami, Florida, has been appointed a master distributor of Kings' coaxial and special connectors, including a unique line of Teflon™ insulated UHF, N and BNC connectors. Kings' Teflon™ connectors are rated from −60 to +165 degrees C at 1000 volts RMS.

The top-of-the-line PL259 is made using a special TR-4™ metal alloy that resists tarnish and enhances solderability. Retail price of the TR-4™ PL259 is $1.59. Other high quality but less expensive connectors are also available.

For more information or a catalog, contact Nemal Electronics, 1325 N.E. 119th Street, North Miami, Florida 33161. RS#302

hf transceiver

ICOM has announced what it calls the most advanced, highest performing hf transceiver with general coverage receiver available to Amateurs today. The IC-751 features ICOM's new CPU, with internal battery memory back-up, provides many advanced features, such as 32 memories with memory storage of mode and frequency, and the scanning capability to cover large segments of the spectrum very slowly, or to scan the memories by selected



mode. The IC-751 provides instantaneous, silent, band selection and has a unique 3-speed tuning system. Other features included in the IC-751 are full break-in keying, passband tuning, notch filter, RIT and XIT with separate readout, fm built-in as standard, a very steep-sided FL44 sideband filter, continuously adjustable noise blanker levels, dual VFO operation, and all mode squelch. An easy-to-read two color fluorescent readout showing the frequency in white and the control functions in red, for low eye fatigue and high visibility in all ambient

✓ 136

NEW products

light conditions is standard. The IC-751 is equipped standard for operation from 12 volts dc, and there is an optional internal ac power supply. The IC-751 has an advanced receiver design that provides true competition-grade performance.

For further information, contact ICOM, 2112 116th Avenue, N.E., Bellevue, Washington 98004. RS#303

improved antenna

Bilal's new Isotron 20 is easier to tune, more versatile in mounting, and covers a greater frequency range than previous models. The new model is omnidirectional and will handle the full legal limit of power. It's adjustable for resonance, weighs approximately two pounds, and measures only 21 inches in length.

The Isotron 20 requires a single coaxial feedline; no tuners or radials are necessary. Its small size makes the Isotron 20 ideal for travel or for use in limited space. The price is $49.95 plus $3.50 for shipping via UPS.

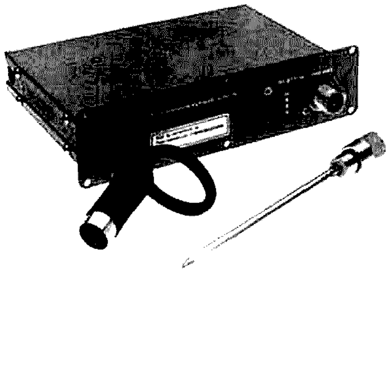
For further information, contact the Bilal Company, Star Route 2, Eucha, Oklahoma 74342. RS#304

vhf converter

Trio-Kenwood has recently announced the release of an optional accessory VHF converter — the VC-10 — to accompany its highly sophisticated R-2000 communications receiver introduced last December. The VC-10 allows the R-2000 to receive signals in the 118-174 MHz range; through the use of microprocessor technology, frequencies in this range may be tuned, displayed, stored in memory, recalled, and scanned, using the R-2000 front panel controls and frequency display.

The R-10 installs easily on the rear panel of the R-2000.

Additional information is available from Trio-Kenwood Communications, 1111 West Walnut Street, Compton, California 90220. RS#305

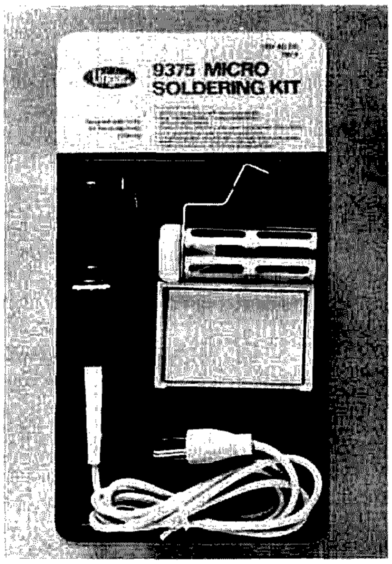


NEW products

microcircuit soldering kit

A kit including soldering iron, tips, and holder for precision microelectronic soldering is now available from the Ungar Division of Eldon Industries, Inc.

The Ungar 9375 Micro-soldering kit includes a three-wire handle, slimmer than those on earlier models, to facilitate close-tolerance soldering. A "Thermo-duric" heating element allows the heating elements to reach working temperature faster, recover more quickly, and use less energy, and last longer than other types of heating elements. Electric leakage, which could ruin microcircuits, is also said to be eliminated.



Three precision tips of differing configurations (needle-point, spade point, and screwdriver) are also included with the kit. Nine additional tips are available.

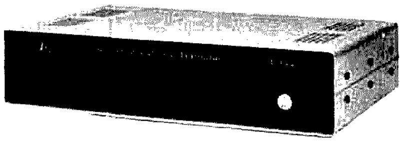
The handle, heater, and tip are all modular, permitting quick replacement or change. The price is $51.75.

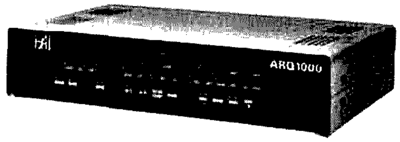
Further information is available from the Ungar Division of Eldon Industries, Inc., 100 West Manville, Compton, California 90220. RS#306

new from HAL

The HAL CT2200 is the successor to the popular CT2100 Communications Terminal. It offers all the features of the CT2100 plus keyboard programming of all eight "brag-tape"

messages, programmable selective call control of the printer output, manual printer on-off control, non-volatile storage of HERE IS, "brag-tape" capability, selective call codes, and new rear panel connections for use with the ARQ1000. While the CT2200 is a new product that replaces the CT2100, a kit (including a new front panel) is available to enable CT2100 owners to update their units to CT2200 capability.





The CWR6750 is a receive-only RTTY and CW demodulator and display generator. The CWR6750 features a built-in 5 inch video display. Operating from +12 Vdc, this compact, portable unit is recommended for RTTY and CW *short-wave listening*.

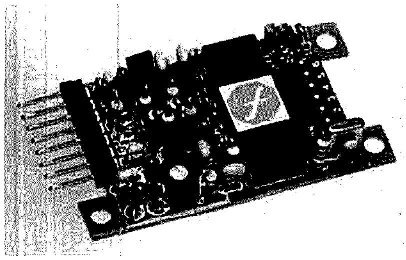
For further details, contact HAL Communications Corp., P.O. Box 365, Urbana, Illinois 61801. RS#307

programmable CTCSS encoder/decoder

Ferritronics has announced the availability of the new P505A CTCSS Encoder/Decoder.

The unit features quartz-accurate frequency synthesis and DIPSWITCH programming to all 37 EIA sub-audible tone assignments.

A choice of plug on or soldered on lead set is offered to suit various mobile and portable applications of the unit, which measures 2.1 × 1.2 × 0.4 inches.



For further details, contact Ferritronics, Inc., 1319 Pine Avenue, Niagara Falls, New York 14301. RS#309

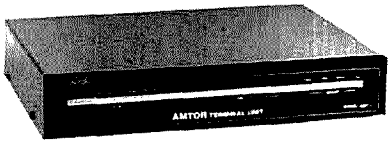
NEW products

AMT-1 terminal unit

The AMT-1 terminal unit contains everything that is needed to convert an Amateur Radio station and personal computer (or ASCII terminal) into a fully operational data communications system. It combines modern circuitry (AFSK modulator/demodulator) with a microprocessor which handles AMTOR data transmission and also translates between AMTOR code and 8 unit, ASCII code.

An ASCII, RS232 interface has been chosen for the AMT-1 because of the extra CONTROL and ESCAPE code flexibility which this allows. Additionally, home computers and data terminals with ASCII interfaces are now available at reasonable prices.

In addition to AMTOR capability, the AMT-1 also transmits and receives standard RTTY and transmits CW (morse code). A fourth "Transparent" or "Direct" mode is available, which



connects the terminal directly to the modem. Using an ASCII terminal, it allows the AMT-1 to transmit and receive ASCII at any suitable Baud rate.

The modem incorporates an active 4-pole receive bandpass filter, feeding into an audio discriminator. It has a performance much higher than that normally offered by Amateur RTTY terminal units. Transmit tones are crystal controlled. Frequency shift is 170 Hz, using the U.S. recommended tone pairs:

2125 Hz (MARK) 2295 Hz (SPACE)

Full status indication is available via LED indicators on the front panel and in addition, a 16 LED tuning indicator has been incorporated. In AMTOR Mode, this acts as a gated frequency analyzer and makes tuning extremely simple.

No switches appear on the front panel of the terminal unit. All control is via ESC and CONTROL functions sent from the terminal or computer.

For complete information, contact Advanced Electronic Applications, Inc., P.O. Box C-2160, Lynnwood, Washington 98036. RS#308

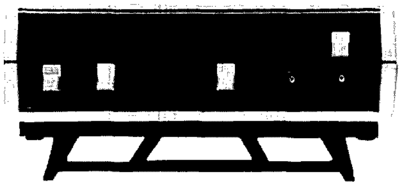
NEW products

frequency counter

Global Specialties' new 1 GHz frequency counter, Model 6002, delivers accurate frequency measurement from 5 Hz to 1 GHz and also measures period from 100 ns to 200 ms. With three selectable resolutions (cycles averaged in the period mode), LED indicators, and simple push-button control, this unit offers uncomplicated operation as well as versatility. A 10 MHz crystal oven oscillator time-base assures ±0.5 ppm (10-40 degrees C), ±1 ppm/year stability.

Two front panel mounted, ac-coupled BNC input connectors provide flexibility in use. The *A* input accepts signals from 5 Hz to 100 MHz with an input impedance of 1 megohm at 20 pF and resolutions of 10 Hz, 1.0 Hz and 0.1 Hz. A switchable lowpass filter provides a 6 dB/octave roll-off at 60 kHz to facilitate audio and ultrasonic measurements. A ×100 multiplication mode is available to speed up measurement and display of frequencies in the 5 Hz to 10 kHz range. The *B* input accepts signals from 80 MHz to 1 GHz with an input impedance of 50 ohms at 10 pF with resolutions of 1 kHz, 100 Hz, and 10 Hz.

The front panel of the 6002 allows ready access to controls. Push-button controls include: Standby/On; Mode; Resolution; A/B input; and Lowpass Filter. The 8-1/2-digit display features leading zero blanking, bright 0.43-inch characters, and a contrast enhancement filter to ensure legibility in ambient light. LED indicators for "Gate Open," "Oven Ready," and "Overflow" provide additional user convenience, and a flip-up leg gives added flexibility for benchtop use.



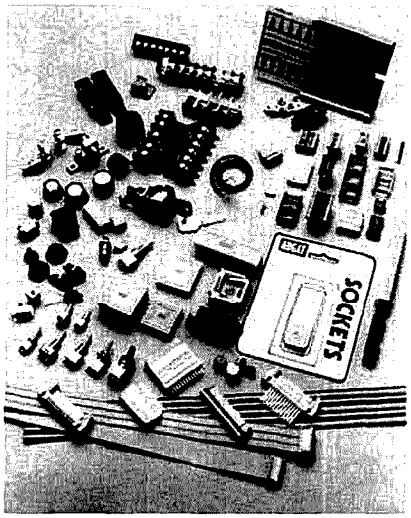
The 6002 can be used for audi-VHF in communications, data processing, process control, rf design, digital design, QC, and maintenance.

For additional information, contact Global Specialties, 70 Fulton Terrace, New Haven, Connecticut 06509-1942. RS#310

NEW products

Augat packaged products

A major industrial manufacturer has announced the nationwide distribution of a complete line of specially-packaged industrial-quality electronic and electromechanical products for consumer use.



More than 600 styles of sockets, interconnection products, ribbon cable, and IDC products and accessories — including Alcoswitch switches, lamps, and knobs are now available to Amateurs.

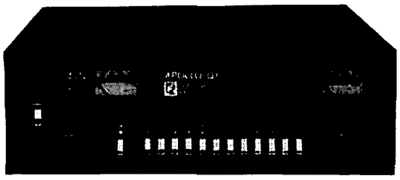
A complete catalog is available. For information, contact Augat, Inc., 89 Forbes Boulevard, Mansfield, Massachusetts 02048. RS#311

satellite receiver

National Microtech, Inc., announces the addition of the Apollo Q-1 satellite receiver and down converter to its product line. The Apollo Q-1 satellite receiver/down converter carries a one-year warranty from National Microtech. It features push-button transponder selection, automatic polarity control, an audio-in signal strength meter display, and a built-in modulator. The Apollo Q-1 is packaged in a woodgrain cabinet with a sleek, black, anodized face plate. A separate down converter with integral LNA power block complete the package.

National Microtech is a leading national dis-

NEW products



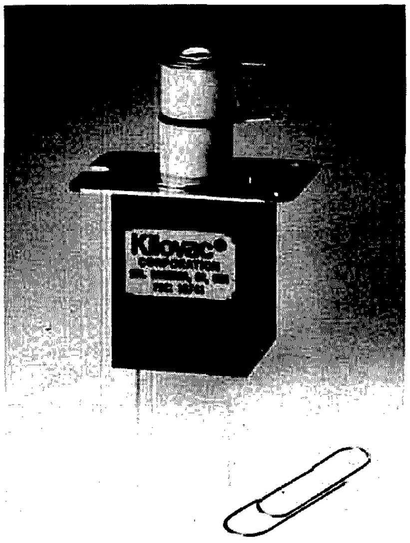
tributor of satellite equipment for the home TVRO and private cable industries. For more information on the Apollo Q-1, contact National Microtech, Inc., P.O. Drawer E, Grenada, Mississippi 38901. RS#312

miniature high-voltage relay series

A low-cost miniature high voltage vacuum relay series with rated operating voltage of 5kV dc and continuous current carrying capability of 15 amperes has been announced by Kilovac Corporation.

The K42A Model has a SPST-NO contact arrangement and is ideal for use in digital antenna couplers where instant frequency hopping capability is desired, as well as in applications traditionally satisfied with open-frame or reed relays.

The new relays are the product of two years of development and testing. They represent



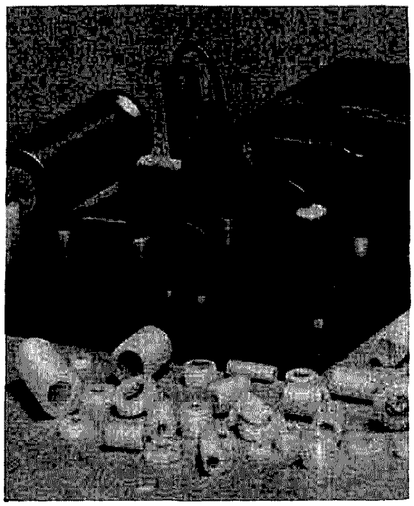
the first in a planned low-cost line of mass-produced, in-stock miniature high-voltage vacuum relays for military and industrial applications. The K42A series relays are designed to meet the requirements of Mil-R-83725. The new ceramic relays are priced from $25 to $30 in quantities of 1000.

Kilovac designs and manufactures electromechanical relays for high voltage, high peak current applications, including medical electronics, ECM, communications, sonar and radar pulse forming networks, antenna switching and coupling, electrostatic coating, dielectric strength testing, laser and x-ray power supplies.

For further information, contact Kilovac Corporation, P.O. Box 4422, Santa Barbara, California 93103. RS#313

universal spacers

Made from natural nylon per MIL-M-20693B, an expanded line of permanent spacers from



Bivar, Inc., gives users an extremely wide selection of tubular spacers for mounting PCB discrete devices in almost any lead pattern and elevation.

Simple in design and extremely easy to use, four basic ID/OD combinations with thicknesses ranging from 0.020 inch through 1.250 inches, in 0.005-inch increments, are available immediately from factory stock or within two weeks. There are 988 standard sizes to choose from, and special heights are also readily available.

The broad selection permits the user to choose the most suitable sizes for strength, elevation, and ease of assembly, as well as to provide for proper filleting and cleaning. Simplified mounting of devices with unusual configurations is easily accomplished.

A typical part, 902-070 (0.32-inch ID, 0.125-inch OD, and 0.070-inch thick), is priced at $10.00/K, in 10K lots. For further information, contact Bivar, Inc., 4 Thomas, Irvine, California 92714. RS#314

self-contained keyer

A new *completely self-contained keyer* from Globalman features variable speed, monitor volume control, automatic or semi-automatic switch, and an on/off switch on the front panel. The other controls such as tone, output keying switch from transistor to relay, external speaker or earphones, battery option input jack, output terminals, fuse, and ac line cord are on the back panel.

The relatively heavy keyer is housed in a crackle-finished steel cabinet designed to remain stationary in use.

An unconditional guarantee covers materials and labor for one year from date of purchase, as long as the keyer is not dissembled or abused. Should malfunction occur, the factory will repair the keyer at no cost to the owner.

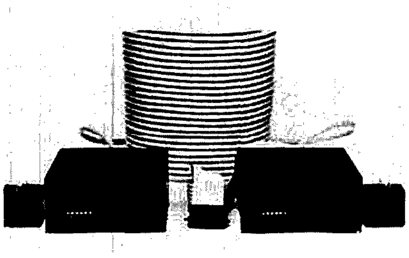
For further information, contact Globalman, Inc., El Toro, California 92630. RS#315

cable stretcher

What do you do when your printer — with 15 feet of cable — is needed 50, 100, or 1000 feet away from your computer?

The new "cable stretcher" from Cronos can help keep your computer and printer in two-way communication without loss of speed or degradation of performance. Model 100LS consists of two heavyweight stretcher boxes, one at each end of the center cable.

The stretcher comes with 25 feet of twisted 15-pair cable (extra lengths may be ordered) up to 1000 feet. Each end terminates in a DB-37P connector. The interfaces weigh about 3 pounds each, use 3 VA of 120 volt 50-60 cycle



power, and will operate in ambient temperatures of 0 to 70 degrees C. The size is 6-1/2 × 6-1/2 × 3-1/4 inches. Six feet of flat cable connecting stretcher to printer are supplied.

For more information, contact Cronos Engineering, Inc., 105 N.W. 43rd Street, Boca Raton, Florida 33431. RS#316

NEW products

electronic grade coolant

A newly formulated electronic circuit/component coolant from Chemtronics is said to make thermal intermittent trouble shooting of electronic circuits components easier than ever. Applied as an aerosol spray, *Freez-It*,™ freezes to −65 degrees F (54 degrees C) instantly. Sprayed copper circuitry frosts and turns white, exposing hairline cracks in PC boards. Rapid chilling of suspected components allows individual components to be isolated for instrument testing. Defective resistors, transistors, capacitors, and other component parts may be quickly and easily identified.

Freez-It can also be used to prevent transformer burnout. If a transformer starts to smoke, spraying the transformer will keep wax, varnish and shellac from running and causing short circuits. It will also freeze adhesives for quick removal.

Other uses include low temperature testing of circuits and equipment, preventing cold solder joints and soldering delicate, heat sensitive components, and as an aid in shrink fit assembly.

The product can be used safely on plastics, rubber, paints, or metal and leaves no residue. It is also non-flammable, non-toxic and odor-

less. Active ingredients are 100 percent pure Freon 12. *Freez-It* has been tested to meet Federal Specification BB-F-1421.

A specially designed 3-way variable control valve controls the spray zone; a pinpoint appli-

cation or wide area coverage can be selected. An extension tube is also supplied for difficult to reach areas.

Freez-It is supplied in 15 and 22 ounce aerosol spray cans.

For further information, contact Chemtronics, Inc., 681 Old Willets Path, Hauppauge, New York 11788. RS#317

clamp-on ammeter adapter

The new Triplett Model 10-N clamp-on AC ammeter adapter is universally adaptable to any digital multimeter with 3/4 inch center banana plugs, 10m ohm input and a 200 mV ac

current range. Said to be ideally suited for in-field, non-interruptive circuit testing, the Model 10-N may be used with a line separator (Triplett Model 101), for single-conductor current measurements on two conductor cables.

Current ranges are 0-20 amps and 20-200 amps with an accuracy of ±3 percent. The range switch may be operated under load with no damage.

The spring-loaded clamp jaws permit simple one-hand operation. Model 10-N is molded from high-impact thermoplastic material to provide years of durable, trouble-free operation.

For further information, contact Triplett Corporation, One Triplett Drive, Bluffton, Ohio 45817. RS#318

new Sinclair catalog

Sinclair Radio Laboratories, Inc., has released the new edition of their product information catalog. The 16-page booklet contains updated technical specifications on a full range of Sinclair's multicouplers, combiners, duplexers, and ferrite accessories. Copies are available from Sinclair Radio Laboratories.

For more information, contact Sinclair Radio Laboratories, Inc., 14614 Grover Street, Suite 210, Omaha, Nebraska 68144. RS#319

NEW products

new Hamtronics® catalog

Hamtronics, Inc., announces publication of their 1983 catalog for the VHF/UHF/OSCAR enthusiast and two-way shops. The thirty-six page, two-color catalog features many new products, including fm repeaters, new VHF and UHF fm receivers, helical resonator preamps and filters, low-noise receiver preamps, and a UHF receiver for listening to the space shuttle. Also included are the popular fm transmitters and power amplifiers, VHF and UHF receiving and transmitting converters, VHF transceivers, and other products.

For a free copy, write to Hamtronics, Inc., 65F Moul Road, Hilton, New York 14468. RS#320

tone signaling products catalog

The new product list from Communications Specialists includes their new direct plug-in CTCSS encoder-decoder boards for many

two-way radios, paging, two-tone sequential, and burst tone devices. For a copy of the list, contact Communications Specialists, Inc., 426 West Taft Avenue, Orange, California 92667. RS#321

short circuits

6-meter amplifier

In the April, 1983, article by W2GN, "6-meter Amplifier," the two (parallel) plate blocking capacitors (**fig. 1**, page 73) are shown incorrectly wired. A correct installation would have them wired between the plate's copper strap and the inductor L as shown correctly done in **fig.** 2 on page 74.

remote control hf operation

The following corrections should be made to K5QY's article, "Remote Control hf Operations" (April, 1983):

The source of the eight-bit I/O card referred to on page 32 is Applied Engineering, Inc., P.O. Box 470301, Dallas, Texas 75247. The price is $62.00.

Line 3775 of K5QY's computer program (page 42) was omitted from the text. Line 3775 should read:

```
3775 IF T = 9 THEN DL = -.0001
```

In **fig. 4**, the center-tapped connection of the two series 0.01 μF telephone line shunt capacitors should only go to ground. The lower telephone line was inadvertently shown grounded.

noise bridge

The 365 pF capacitor required for construction of K2BT's rf impedance bridge (March, 1983) can be obtained from Radiokit, Box 411H, Greenville, New Hampshire 03048, or from Mouser Electronics, 11433 Woodside Avenue, Santee, California 92071.

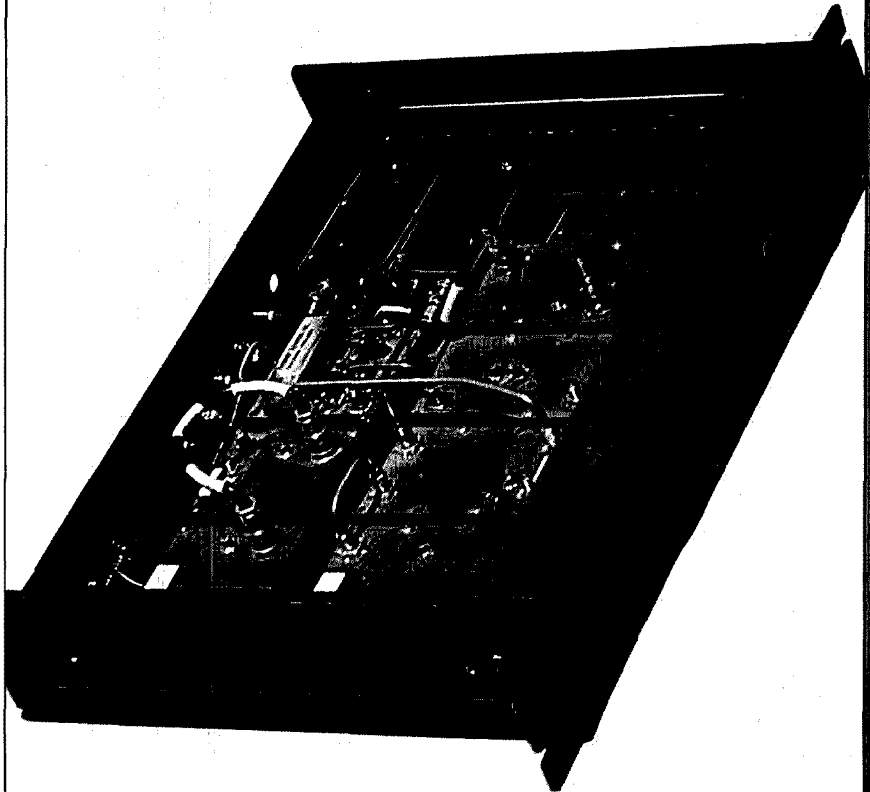
keyer interface

In the February, 1983, article entitled "Low-Power Keyer and Interface," the following CMOS chips were used: type 4023 for U1, 4013 for U2, and 4001 for U3. Also, power (V+) is applied to pin 14 and GND to pin 7 of each chip.

NBC's Roy Neal, K6DUE: Working W5LFL from Space

ham radio magazine

- rf synthesizers
- fail-safe power
- locating orbiting satellites
- effective ground techniques
- local weather radar



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focus on communications technology

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contents

REFLECTIONS

a personal phone call

If I could pick up the phone and call you right now, I'd ask each one of you personally how *you* feel about *ham radio magazine.*

Something akin to this actually happened at Dayton this year, when more than 400 readers took the time to express their views, needs, and concerns to me at our booth. But this number — impressive as it is — represents less than one percent of our subscribers . . . and we need to hear from the other ninety-nine percent.

Since I can't call each one of you personally, I'm doing the next best thing: asking you to take advantage of the survey form printed on this page and the next, and let us know, very clearly, how you feel about *ham radio.*

I see this magazine (and all magazines, for that matter) as a conduit of thoughts — a communications channel. Editors are in a privileged situation. Information arrives from all directions, and can be directed, stopped or modified. Add to that an editor's own personal interests and inclinations and one can see that the different directions a magazine can take are many. But my desire is to provide material that *you* want to read and use.

Because Radio Amateurs are such a diverse group, I fully expect to hear completely opposing viewpoints expressed with fervent conviction. The majority will rule, of course. But the minority will be heard. And we'll do what we can to try to meet their needs, too, without compromising our primary mission, which is to provide the Amateur community with the most useful and informative technical material in the Amateur Radio publishing field.

We all have vested interests in this wonderful hobby; each and every one of us, at one time or another, has expressed our views privately or in the presence of large groups. All I'm asking is for you to do the same now in responding to the questions that follow.

Many of the questions, you'll notice, have been asked before in other surveys. We have to ask them again because an accurate, up-to-date reader profile helps us make the short and long-term decisions that result in a magazine that suits your interests as closely as possible. Some questions ask you to rate the magazine by itself and in comparison to others in the field. (Besides being curious, we at *ham radio* prefer to work with fact rather than fiction.) Other questions attempt to determine, very specifically, your present and future needs. We're only human. Maybe we've left out an area or two you would like to see discussed. Write it down!

What is at stake are your needs. Correspondence and communications — dozens of letters and phone calls each day — coupled with our best intentions and understanding indicate that *ham radio* should continue to provide a technically sophisticated subject menu. With your help we'll be able to vary your diet, make it fuller and more satisfying. Don't stop now — take pen in hand and as neatly as possible (please think of our poor compilers) respond to the survey questions. *All* of your opinions are wanted and needed. Thank you.

Rich Rosen, K2RR
Editor-in-Chief

1983 reader survey

Please reply to the questions below and return this page (or a photocopy) to READER SURVEY, Ham Radio Magazine, Greenville, New Hampshire 03048 by November 1, 1983.

ABOUT YOUR AMATEUR INTERESTS

a. Class of license held__________

Check here if no license held__________

b. Number of years licensed__________

c. In which modes do you operate? (Check all that apply)

	Exclusively	More than 50%	Less than 50%	Never
CW				
SSB/AM				
A5				
RTTY				
FM				
Packet				
Other				

d. What frequency ranges do you work? (Check all that apply.)

	Exclusively	More than 50%	Less than 50%	Never
l-f (1750 meters)				
m-f (160 meters)				
h-f				
vhf				
uhf				
microwave				

e. How much of your equipment do you build?

100%____ 75%____ 50%____ 25%____

Less than 25%____ None____

f. When you build equipment, do you prefer to build:
from a kit____ from "scratch"____

g. What was the most complex piece of equipment you've built in the last 5 years? ____________

h. How many hours a week do you spend on Amateur Radio operation and activities? ____________

i. About how much do you spend each year on Amateur equipment?

$100-$500____ $500-$1000____ $1000-$2000____
$2000-$4000____ $4000+____

j. From what source do you obtain your books about Amateur Radio?
dealer____ Ham Radio's Bookstore____
other mail order____ at shows____

k. Do you own a personal computer?
Yes_____ No_____
If "yes," what kind?____________
If "no" do you:
Plan to purchase within one year ____________
Do not plan to purchase ____________

ABOUT HAM RADIO MAGAZINE

a. From the following list, please choose the five areas of interest you would most like to see addressed in *ham radio* magazine on a regular basis. Enter your choices, in order of preference (favorite first) in the spaces below:

antennas
awards
book reviews
computers and Amateur Radio
construction (simple)
construction (intermediate)
construction (advanced)
contests
DX
FCC news
future technology
interviews
international Amateur Radio journals
license upgrading
new product announcements
News in Amateur Radio and current events
opinion
operating events
product reviews
Questions and Answers
radio history
receivers
satellites
social events
speech synthesis
surplus (conversions)
test equipment (to build)
test equipment (to buy)
test equipment (to use)
theory (simple)
theory (intermediate)
theory (advanced)
other (please specify)

1. ____________
2. ____________
3. ____________
4. ____________
5. ____________

b. To which Amateur Radio publications do you subscribe?
*ham radio*________ QST________ CQ________ 73________ Other____________

c. Of the publications listed above, which one suits your needs best? ____________

d. In general, how do you feel about *ham radio* magazine?
best in the field________ better than most________ satisfactory________ unsatisfactory________

e. In regard to technical content only, do you feel *ham radio* magazine is:
about right________ too technical________ not technical enough________

f. In regard to clarity and readability, do you find *ham radio* magazine:
well-designed and edited, easy to read________ satisfactory________ needlessly complex________

g. In general, do you feel that the articles in *ham radio* are:
about right________ too long________ too short________

h. Have you ever purchased Amateur equipment as a result of reading about it in *ham radio* advertising?
Yes________ No________

i. Have you ever purchased Amateur equipment as a result of reading about it in *ham radio* editorial content?
Yes________ No________

j. Have you ever purchased *commercial or industrial* equipment as a result of reading about it in *ham radio* advertising?
Yes________ No________

k. Would you be interested in reading about current developments in *commercial or industrial* radio equipment in the pages of *ham radio*? Yes________ No________

l. Looking back over the past three months (June, July, August) what one article or feature in *ham radio* did you *most* enjoy? ____________
Which article or feature did you *least* enjoy? ____________

m. If you could tell the editors of *ham radio* how the magazine could be improved, what would you say?
____

ABOUT YOU

a. Your age:________
b. Occupation:________
c. Income Range (check one): Below $7500______ $7500-$12,999______ $13,000-$19,999______
$20,000-$29,999______ $30,000-$49,999______ $50,000+______

Thank you for taking the time to complete this survey.

presstop de W9JUV

1500 WATTS PEAK OUTPUT WILL BE THE NEW LEGAL LIMIT for U.S. Amateurs effective August 29. Acting on PR Docket 82-624, the Commissioners agreed on the new output limits for both SSB and CW July 18. 200 watts out will be the new limit for Novice bands, but 30 meters is still 250 watts input. In addition, AM users will be "grandfathered" use of 1000 watts DC carrier input until June, 1990, as the new limit represents a power decrease for them. At the same time, the FCC deleted the requirement for each Amateur to have instrumentation to assure compliance with the rules on power limits—but individual operators better be sure they're operating within the limits should the FCC decide to inspect their stations!

Next Major New Issue Likely To Appear On The Commission's Agenda is "broadcasting," specifically the endless feuding that occurs between participants in some Amateur nets and individual operators. Expect the subject to surface as a Notice of Inquiry this fall.

OSCAR 10'S NOT-QUITE-AS-PLANNED INITIAL ORBIT apparently resulted from an after-separation collision with the launch vehicle's third stage. Supporting indications include telemetry data from vibration sensors and an apparent change in the radiation pattern of OSCAR 10's 2-meter antenna, a change that could have resulted from collision damage.

The Important Damage Was To The Kick Motor, Which Fired OK for the first burn but did not respond to final burn commands July 25. Though this leaves the orbit significantly off nominal, its final 3951 by 35505 km elliptical path will still provide long openings for users throughout the world. The fact that OSCAR 10 is up and fully operational reflects tremendous credit on its designers, builders and controllers. Just getting a satellite as sophisticated as OSCAR 10 up and operating is a challenge to the most experienced professionals, and for a group of "Amateurs" to accomplish the feat despite a potentially catastrophic accident is almost beyond belief! Congratulations to AMSAT's "Amateurs!"

OSCAR 10's Transponders Should Be Turned On for users by sometime in mid August.

SUPPORT FOR CONTINUING CW LICENSE REQUIREMENTS HAS COME from an unexpected source, the U.S. Air Force, which has advised the FCC it's reinstating Morse proficiency for its radio operator training course. This probably means that a no-code Amateur licensee would not be eligible for Air Force MARS without demonstrating CW ability.

Senator Barry Goldwater Has Strongly Restated His Opposition to "No-Code" in a "Dear Mark" letter to FCC Chairman Mark Fowler. In it he says "I completely support (the ARRL's) Comments" against no-code, and says he believes it would be "a grave mistake to remove the code as a license requirement...." The QCWA also filed strong comments against no-code.

A Decision On No-Code Appears Likely By Year's End, but at this time no one seems to want to make odds as to which way it will go.

"AMATEUR RADIO'S NEWEST FRONTIER," THE NEW AMATEUR RADIO VIDEOTAPE production, is well on its way to release. NBC's Roy Neal, K6DUE, and Westlink's WA6ITF did the taping in mid-July at the ARRL, FCC, AMSAT's Goddard facilities, and with NASA in both Huntsville and Houston. Editing and dubbing were completed July 24, and the new 16-minute Amateur Radio promotional piece should be available about September 1 on VHS and 3/4-inch tape.

THE FCC'S NEW "NO-MAILBACK" NOVICE EXAMS MUST CONSIST of one, two, or three questions from each of 20 FCC-provided groups of 10 questions each. The answers are not included, perhaps because the examiner, as a higher class licensee, should already know what the correct answer--or sometimes answers--is. In addition, it might also give a better opportunity for the examiner to determine if the applicant really knows what he should or is simply parroting back answers from a "study guide."

A Copy Of The FCC's List Of 200 Questions and details on how to administer the new Novice exam are available from ham radio for a business size (#10) SASE.

W5LFL's OPERATION FROM THE SPACE SHUTTLE IS STILL ON SCHEDULE for the end of September. The ARRL is still trying to arrange for a "900" information line to be operational about a week before launch. Information on that line is to be updated frequently, particularly if it pertains to W5LFL's on-the-air activities.

145.55 MHz Should Be W5LFL's Prime Transmit Frequency, with several alternates to be announced shortly before launch along with about 20 receive channels.

"SUPER STRICT" INTERPRETATIONS OF THE FCC'S "BUSINESS COMMUNICATIONS" action that was announced in last month's Presstop were not what the Commission intended. The FCC's action actually changed little in the rules, but was simply meant to be a reorganization and clarification of business communications-related items already there.

Public Service Type Operations Such As Parade or walkathon coordination and the Eye Bank Net are still permitted. Amateur-related operations such as Hamfest "Talk-Ins" and even on-the-air swap nets are also still legitimate. In its release, the FCC stated: "The Order was not intended to impose any new restrictions or to cut back on what Amateurs have legitimately been doing all along. What was intended was to alert the Amateur community to the fact that the Amateur Radio Service should not be used in lieu of other radio services for the transmission of business messages."

Catalysts For The FCC's Action Were Two Recent West Coast Episodes, one in which a "citizens' posse" began using a local repeater as an auxiliary police communications channel. In the other a government facility wanted its employees with Amateur licenses to patrol its fences and gates while using 2 meters for coordination.

PRIVATE RADIO BUREAU CHIEF JIM McKINNEY HAS RESIGNED to become Chief of the FCC's Mass Media Bureau. Jim, who's long been a knowledgeable supporter of the Amateur Service, is replaced by Private Radio Bureau Deputy Chief Bob Foosaner.

comments

PEP talk

Dear HR:

The article by VK3AFQ (*ham radio,* June, 1983), demonstrates one of the fallacies of defining the legal power limit in terms of PEP output. The metering device described, like the commercially available in-line "wattmeters," is designed to work with a 50-ohm non-reactive load. Not all Amateur stations are equipped with suitable transmatches to arrive at a perfectly matched line at all frequencies within a band — let alone homebrew rigs designed to work into balanced feedline or oddball impedances.

Nor is the statement by Rich Rosen, Editor-in-Chief, that the FCC proposal would increase our power to 1500 PEP output, strictly true. Under present rules there is no limit to transmitted PEP, but practically, the limit is at least 3000 watts (one killowatt input, plate-modulated a-m phone). The power limit for CW, fm, RTTY, and SSTV would be doubled, while the limit for SSB would remain essentially unchanged.

Actually, PEP is useful as an equipment design parameter, but is quite irrelevant to the purpose behind the legal power limit. It is mean, or average, output power that determines the effective field strength of a signal, not the power developed on occasional voice peaks. I believe that if the FCC goes ahead and replaces the relatively simple input power limit with something as complicated to measure as PEP output, the practical effect will be the elimination of a power limit altogether. Many otherwise law-abiding Amateurs will rationalize illegal power by claiming to lack the facilities to make the necessary measurements to comply voluntarily under the new rules. The FCC simply does not have the facilities to routinely inspect Amateur Radio stations for transmitter power.*

Donald Chester, K4KYV
Woodlawn, Tennessee

*It's now official — see *presstop* for details.

simpler "panadaptor"

Dear HR:

I've been an *hr* subscriber since April, 1972, and I can't remember an issue in which I've found more first class technical articles than your February, 1983, number. Congratulations!

Rick Ferranti's "Design Notes on a Panoramic Adapter/Spectrum Analyzer" was particularly interesting. As someone who has worked with a similar homebrew project, I'd like to offer a couple of additional notes.

If you have a receiver with a narrowband CW filter and you don't mind poking a few holes in the cabinet, it may be possible to simplify the panoramic adaptor and reduce the parts count by using the receiver's internal mixer, i-f amplifier chain, and filter. This technique involves substituting an external swept oscillator for the tunable local oscillator of the receiver and substituting the detector and video amplifier of the spectrum analyzer (Ferranti's fig. 11) for the receiver's product detector and audio chain. The swept oscillator is connected to the LO port of the receiver's internal mixer stage in place of the internal tuneable LO; and the i-f output is picked off after, rather than before, the receiver's high selectivity filter.

Before this arrangement, the external circuits which comprise the panoramic adapter consist only of: (1) a sweep oscillator, (2) a new "local oscillator" for the receiver having the same tuning range as the receiver's own LO but which can be swept in frequency by the sawtooth output of the sweep oscillator, (3) the detector/video amplifier, and (4) the oscilloscope used to display the signal. For this circuit to function properly, both the tuneable local oscillator and the AGC bus of the receiver must be disabled. The receiver's mixer and i-f amplifiers should have wide dynamic range for best results. Sweep speeds ranging from 50 to 2000 milliseconds are desirable to avoid ringing effects, and these are practical if a long persistence (P7) CRT is used for the display. Finally, it should be noted that this technique is suitable only for receivers that have the same tuneable LO tuning range on each band; but this is true of most modern receivers that use a crystal-controlled first converter stage.

Ferranti's design has the advantage of making it possible to sweep across a wide band of frequencies, regardless of the i-f and LO frequencies of the receiver with which it is used. Using the scheme I recommend, this performance is harder to obtain.

If the highest LO frequency is less than 3 or 4 MHz (this would be true of most "Collins type" receivers using 455 or 500 kHz mechanical filters), there is a simple solution, using the Exar XR-205 "Monolithic Waveform Generator" chip as a swept LO. This chip has a guaranteed sine wave output of at least 2 MHz, and individual chips will run as high as 4 MHz. The frequency can be swept over a wide range, and the sweep is extremely linear with applied voltage.

Where the operating frequency of the LO is higher but still too low to permit the oscillator to be swept by a voltage-variable capacitor of reasonable size, some form of heterodyne scheme is a must. Ferranti's 32.82 MHz VCO (**fig. 5**) can be used, mixed with the output of a fixed crystal oscillator. After passing the difference frequency through a lowpass filter, the resultant would be injected at the LO port of the receiver's mixer stage. Even with this added complication, the recommended circuit will have a far lower parts count than Ferranti's double conversion technique, and there should be fewer critical adjustments.

Miles B. Anderson, K2CBY
Sag Harbor, New York

more 10-meter beacons

Dear HR:

A 10-meter beacon, using callsign KA1YE/B, transmits from Connecticut on 28.284 MHz. It sends a 30-second carrier followed by "KA1YE/B SE CT" and operates continuously at a 4 watt output level to a vertical antenna.

The May, 1983, issue of *Practical Wireless* lists 19 worldwide beacons. A list of active beacons is maintained by Willi, HB9AVE. He would appreciate information from any groups or individuals either operating or planning to operate a 10-meter beacon. There is a limited amount of space allocated for beacon operation on 10 meters here in the States, so some prior listening and coordination is highly recommended before activating a beacon.

I am willing to act as a clearinghouse for U.S. beacon information and activities. If anyone is planning to put a 10-meter beacon on the air, or knows about any other beacons active or planned for the U.S., I would appreciate hearing about it.

W. Keith Hibbert, KA1YE
25 Hillcrest Road
Niantic, Connecticut 06357

✓ 156

linear translators

Narrowband technology and linear systems show advantages over conventional fm designs

Narrowband techniques on VHF, namely SSB and CW, have attracted increasing interest among Amateurs over the past few years. While fm and repeaters provide excellent service to thousands of hams, it is apparent in heavily-populated areas that we cannot go on forever with modes requiring 30 to 40 kHz per QSO (the spectrum needed for a repeater input and output). Several solutions to this problem have been proposed, including single-sideband fm, spread-spectrum techniques, and, more recently, ACSB.[1] With a 3:1 to 4:1 advantage of SSB over fm a 3 MHz repeater pair could support as many as 300 repeater channels using SSB-based techniques. Use of these narrower modes places a new requirement on the "repeatering" techniques which are nicely met by LTs or linear translators.

what is a linear translator?

A linear translator (LT) is a multi-mode repeater — with several important differences from fm repeaters.[a] The basic difference between a translator and a repeater is in detection (fig. 1). A repeater detects an incoming signal and then uses the resulting audio (or video) to remodulate a transmitter. A translator merely converts the signal to a convenient i-f frequency, amplifies it, and re-converts it to the desired output frequency without ever detecting the signal; that is, the entire system operates at rf frequencies rather than remodulating a transmitter with detected audio or video. All rf stages in a *linear* translator must operate in their linear region. No limiting or logarithmic processing can be applied. In this way it differs from the fm broadcasting translators used in fringe areas to improve weak signal reception.

A repeater can receive only one frequency and one mode at a time, and retransmit what it hears via only one mode. A translator does basically the same thing but it does it at rf, so that if its i-fs are linear, any mode or signal inserted at the input is reproduced at the output. Even multiple signals at the input will be reproduced at the output.

By James Eagleson, WB6JNN, 280 Manfre Road, Watsonville, California 95076

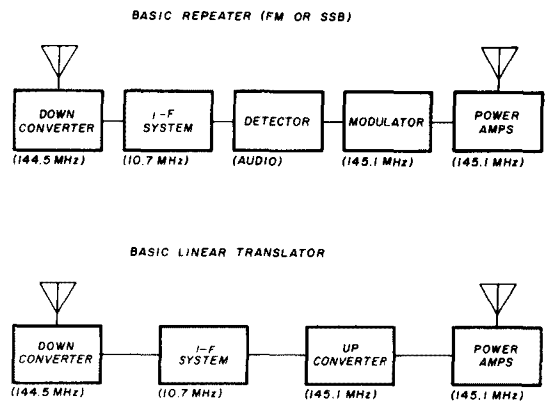


fig. 1. Block diagrams of a repeater and a linear translator.

history of LTs

In early 1980, two linear translators were placed into operation in the San Francisco Bay area.

One machine, built by Narrow Band Communicators, Inc. (NBC, Inc.),[b] is an in-band translator using a standard 600 kHz split. It incorporates two independently AGC'd 10-kHz-wide i-fs and can support as many as four SSB QSOs with minimal interference or two, low-deviation fm QSOs.

The other LT, built by Project OSCAR in cooperation with the Bay Area Two Twenty Group (BATTG),[c] uses an input of 1296.3 MHz and an output of 1269.3 MHz. Its i-f bandwidth is 30 kHz. Its proposed uses as an SDLT (Satellite Development Linear Translator) are:

... to act as a regional "repeater."
... to provide net and bulletin services.
... to provide a source of 1269 MHz signal for OSCAR, Phase III, mode L (24 cm/436 MHz) operation.
... to provide multiple-channel capability for SSTV, Packet Radio and RTTY users.
... to provide experimental data needed to build satellite interties, i.e. interlink translators between satellites.
... to provide an economical means of accessing the proposed commercial geosynchronous satellite with an amateur channel add-on.
... to provide intra/interstate network interties via unused LT channels.

comparing linear translators and fm repeaters

A linear translator provides all the same functions normally provided by an fm repeater. These include extending the coverage of mobile transceivers; extending the coverage of stations with low power, poor locations, or limited antennas; providing a fixed monitoring frequency; and making available a known frequency, power level, and location useful for station evaluation. Additionally, a translator can provide multi-mode operation; multi-station operation; CW (A1); and telemetry, remote sensing, and codestore (mailbox or bulletin) capabilities, all independent of main channel use. With narrowband techniques, both voice and data channels can use an LT simultaneously with only moderate interference.

Although the basic difference between repeaters and LTs is in detection, some operational differences also exist. First, a linear translator does not produce a squelch tail. Also any frequency variations by the incoming signals will also appear at the output — with the same variation. While the noise output of a linear translator is 10-20 dB less than that of an unsquelched fm repeater, its noise floor may be detected by stations close in unless some form of squelch or COR system is used. Another difference is that fringe-area stations may not be able to hear other fringe-area stations, even though all stations local to the LT will easily hear and be able to work both (**fig. 2**).

LT design approaches

SSB translator. Sensitivity: 6-9 dB improvement

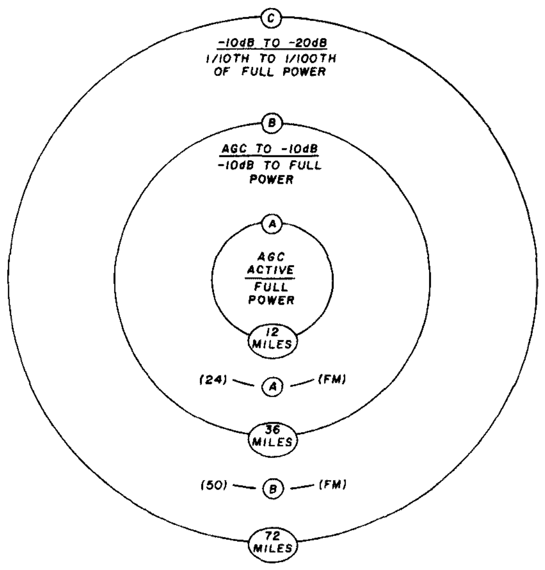


fig. 2. Linear translator coverage diagram. All stations in Region A can hear and work all stations using the LT; stations in Region B can work and hear all but Region C stations; stations in Region C can only work and be heard by Region A stations.

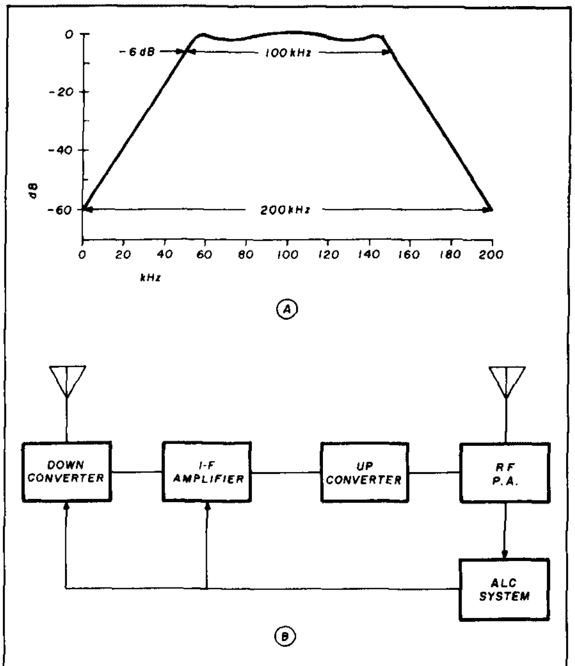


fig. 3. Shared channel block diagram and spectrum requirements.

over an fm repeater;(d) Power output: 200 watts PEP at altitudes limited to 100-watt carrier power on fm; Range: advantage is 1.4:1 over equivalent fm repeater. Using signal-to-noise-ratio enhancement techniques such as ACSB can make this as much as 2:1.(e)

Shared-channel translator. In this system, all stations share one "channel" and derived ALC or AGC is a function of the output of the entire passband (see **fig. 3**). While this technique has the advantage of simplicity and works reasonably well for satellite translators, in which there are only small relative power or range differences between stations, the system is not well suited for land-based use. Several problems exist when using the shared-channel translator with terrestrial LTs that do not occur with multi-i-f LTs.

Selectivity is one of the problems. A 20-kHz i-f filter with a 2:1 shape factor is 40 kHz wide at its −60-dB points. On the other hand, two side-by-side 10-kHz filters are 30 kHz wide at the −60 dB points, since each filter is narrower to start with (**fig. 4**).

Strong signals can desensitize the gain stages in this system and a form of squelch or COR (carrier-operated relay) is needed to reduce or eliminate the LT's noise output when it is not in use. With the single 20-kHz-passband kind of system, the sensitivity of this COR is 3 dB worse than that of a system using two 10 kHz i-fs.(f)

The ARTOB LT. A third method, suggested by Italian ARTOB experimenters,[2] uses multiple i-fs for independent AGC control of smaller portions of a satellite's total passband. This prevents any one station from creating interference on other segments of the passband. The Italian system uses three 30-kHz i-fs yielding a total passband width of 90 kHz. The in-band, 2-meter NBC translator uses a slightly different approach, with two 10-kHz i-fs plus one 3-kHz "single channel" i-f. AMSAT Canada, in their SYNCART proposal,(g) suggests using one 100-kHz general-purpose i-f flanked by two 10-kHz, special-service i-fs (one for data and one for bulletin, net, and special uses). All of these systems were developed about the same time to meet specific requirements.

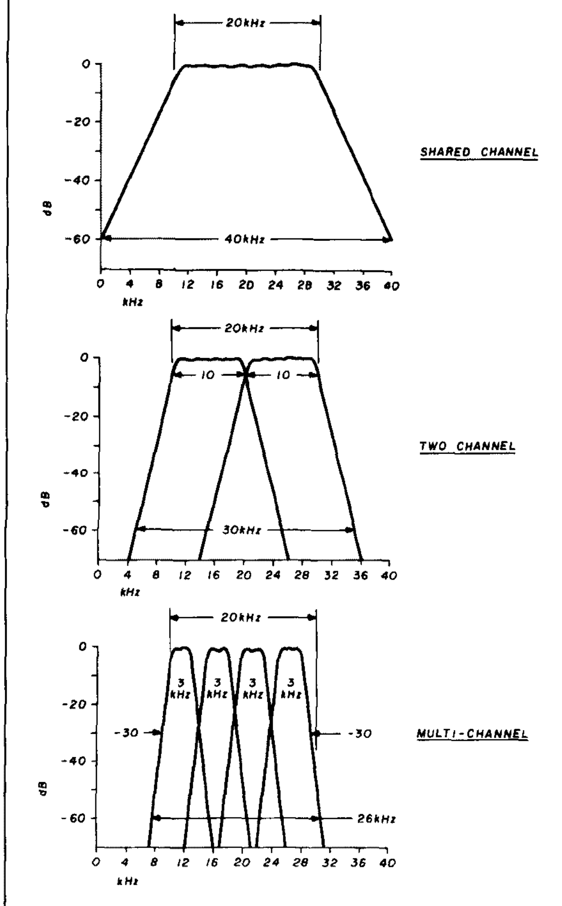


fig. 4. Spectrum allocations for a shared channel, two channel and multi-channel system.

advantages of multiple i-fs

With the use of multiple i-fs, any i-f can be turned on or off at any time to help control interference, reduce noise from unused channels, or to allow re-use of the edge of one LT by the edge channel of an adjacent one. Out-of-channel noise and i-f-caused distortion products can be more closely controlled by the use of multiple, narrow i-fs. A 20-kHz-wide LT using five 3-kHz-wide i-fs has a −60 dB bandwidth of 24 kHz; by contrast, a single i-f system has a −60 dB width of 40 kHz.

Under low-use conditions, when the number of stations is fewer than ten (the normal situation on a 20-kHz-wide LT), better control of AGC pumping by strong stations can be maintained. With a multiple 3 kHz channelized LT all stations can provide full output regardless of the ratio of the strongest to weakest station. Intermodulation-related interference between the various channels can still occur, but even this is improved by the multiple i-f technique.

A multiple i-f channel system can dedicate specific channels for special uses. SYNCART has a hard-limiting fm-AFSK (ASCII) 15 kHz channel for data transmission, a general-purpose, 100 kHz-wide channel for operation similar to present OSCAR satellites, a special 100-kHz i-f for RTTY, SSTV, and ASCII narrowband techniques, and a reserved 15-kHz i-f for educational, bulletin, net, and emergency uses (**fig. 5**). Terrestrial LTs could have 1-kHz i-fs for CW/FSK, 3-kHz i-fs for SSB, 10-kHz i-fs for fm or shared SSB, or some combination providing multiple uses.

system design

Design considerations for a linear translator are somewhat different from those for an fm repeater.[h] For one thing, many integrated circuits, such as the CA 3089 or MC 1357, are not available for linear receivers.

The general scarcity of knowledge about LTs is reflected in the lack of surplus or commercial linear translators, the unavailability of step-by-step LT repeater handbooks, and the absence of experienced commercial LT users. Even AMSAT, probably the most experienced developer of LTs for the OSCAR satellites, has not published much on their design and construction.

building a linear translator

The power amplifier. Multi-station LT operation requires better than average linearity in the PA stages, with greater than 40 dB third-order IM distortion the goal. A maximum output of 100 watts from the amplifier, using two 80-watt stages, should produce an amplifier capable of meeting the design goal. The

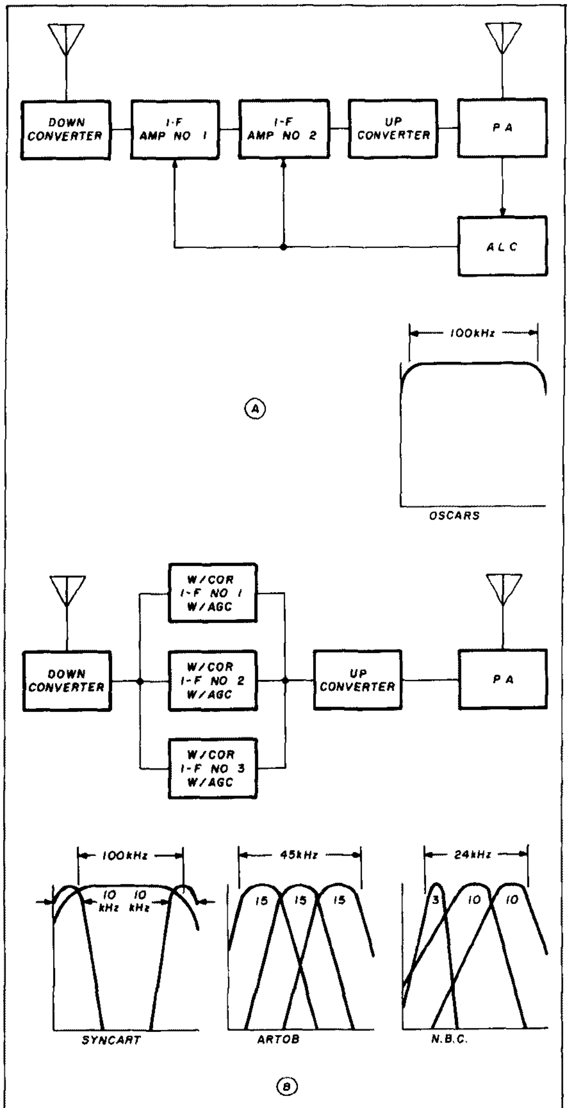


fig. 5. OSCAR transponder and ARTOB, SYNCART, and NBC linear translator block diagrams indicating i-f channel spectrum allocations.

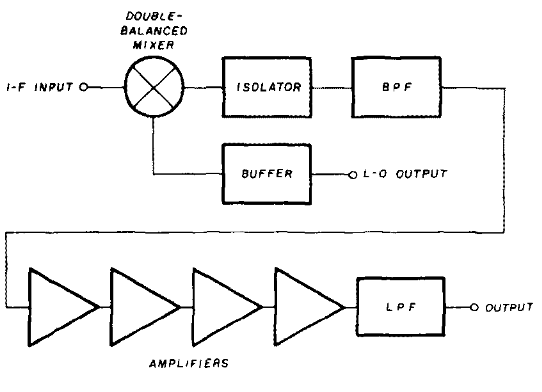
addition of a 3 dB or higher gain antenna brings the ERP level up to 200 watts.

Transmit chain. An output power level of 100 watts helps define the gain required between the transmit up converter and the output port (**fig. 6**). Using a high-level, double-balanced, diode-ring mixer — such as the Mini-Circuits SRA-1H, the Cimmaron CM-1H, or equivalent — IM_3 products should be 40 dB below each output tone at an input signal level of 0 dBm PEP. This translates to an output level of

approximately −7 dBm PEP from the mixer (normal losses) with another 1 dB lost in the image bandpass filter. The resultant input to our transmit PA amplifiers is −8 dBm PEP. An output level of +50 dBm requires +50 − (−8) or 58 dB gain for achieving our goal. **Fig. 7** can be used to work with different power levels.

I do not recommend using lower cost double balanced mixers for the upconverter. A 10 to 25 watt system might be able to make use of them, but the spurious-free output of the high-level mixers is easier to filter than that of the "standard level" units. The addition of a local oscillator buffer helps drive the high-level mixer LO ports to +17 dBm. If standard level mixers are used, all ports must be terminated in 50 ohms at all rf, i-f, LO, harmonic, and image frequencies. Not doing so can affect linearity and spurious content significantly.

The front end. Next in importance to the transmit chain is the front end of the LT. Too much gain can produce intermodulation products, desensitization, crossmodulation products, and other related problems. Too little gain reduces sensitivity for weak-signal reception (poor noise figure) and makes the choice of i-f gain more critical. The best choice for a mixer for the receive converter (down converter) is the double-balanced, diode-ring mixer. Although its inherent loss worsens the front end gain, other kinds of mixers (especially the bipolar) provide gain at the expense of poorer noise figure performance (10 dB typical versus 6.5 dB for a double balanced mixer), poorer IM, desense, or crossmod, or poorer port-to-port isolation.[3] The best compromise seems to be a good low-noise preamp followed by a good double balanced mixer feeding an i-f amplifier with a 2-4 dB noise figure.



recommended active devices

power level	device
100 mW	2N4427, 2N5109
1 W	2N5590, 2N6080
10 W	2N6081, MRF 239, MRF 240
25 W	MRF 240, B-40-12
50 W	MRF 240, B-40-12
100 W	4X 250 (tube)

fig. 6. Block diagram of a linear translator transmit chain. Recommended active devices are listed for each power level.

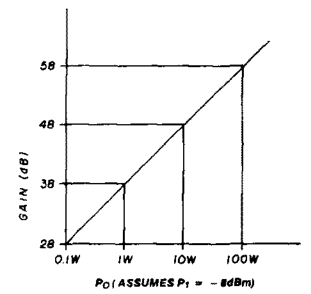


fig. 7. Chart detailing PA gain requirements versus desired output power levels.

The preamp's gain should not exceed 25-30 dB, and it should have a noise figure of less than 2.5 dB. This is easily obtained with single-stage GaAs FETs with noise figures under 1.0 dB. The J-310 and U-310 JFETs can also be used in two-stage preamps yielding a 1.8-2.2 dB noise figure with 25 dB gain. Another good choice is the CP-640 or 643. These units can give reasonable noise figure and excellent high-level performance at higher voltages. The older FETs (such as the MPF-102 and 40673) should be avoided because they do not provide the IMD, crossmod, and desense performance required for LT service.

A word of caution concerning GaAs FETs. Though they provide excellent performance with less than 0.5 dB noise figure and better than 100 milliwatts 1 dB compression point performance on 2 meters, they can be destroyed by improper handling. Oscillations resulting from improper terminations, insertion into systems not having adequate TX/RX isolation in the duplexer system, and static discharges (even when the device is not turned on) can all cause failure. Careful handling is necessary.[4]

Noise floor output. The noise floor is the amplified output of the noise generated in the front end and i-f amplifier stages. In an fm repeater this noise is equal to the output of any station received by the system but it is squelched off when no signal is present and masked by signals when they are there. In an LT, the noise output depends on the gain of the system, overall noise figure, and the amount of external noise amplified by the system. Since it's not desirable for a repeater to transmit constant noise, some control is needed.

For high-altitude, balloon-borne LTs like the ARTOB experiments, or for orbiting LTs such as the OSCARs and RS, no COR or squelch is required.

Everyone is far enough from the LTs so that by setting the noise output at about 30 dB below the maximum LT output level no one complains about the passband noise — even though it is transmitted continuously.

The terrestrial situation is different. Stations may be within one or one hundred miles of an LT, and a 40-dB stronger signal is received at the closer location. At this close station, the noise output of the LT will be heard quite clearly. The answer, of course, is squelch or COR. For the NBC machine we chose to operate the PA systems using COR. Single-channel or multiple-channel systems could use squelch on each channel.

Noise floor versus full output level. A 30-dB full-output to noise-output ratio has been adopted by the ARTOB and OSCAR LTs. Will that suffice for a terrestrial LT? There are several arguments for and against it.

The argument for it is that, almost universally, the typical ham operator begins to feel that a signal is strong when it approaches a 30 dB signal-to-noise ratio (SNR). While we rarely obtain 30 dB SNR on the hf bands (at least signal-to-*interference* levels are more typically in the 5-15 dB range), if there is only noise behind the signal, 25-30 dB SNR is where most of us would start calling it S-8 or S-9.

On the other hand, weaker stations using a system set with the AGC 30 dB above the noise floor are clearly at a disadvantage. A signal arriving at 10-15 dB above the noise floor, for example, would have an output 15-20 dB below the typical strong signal. This means that fringe area stations could access the LT and work stations local to the machine, but would not be heard by another station in their own area or in some other fringe area of the LT.

It is my opinion that an LT using COR or squelch or both can provide more gain so that weaker signals will be more intelligible. The additional noise will be squelched off when the LT is not in use and stronger stations will reduce the gain through AGC action thus reducing the background noise, with a subsequent improvement in SNR. This can be especially helpful in single- or multi-channel systems where AGC pumping of one signal by another signal is less of a factor.

Recommendations. *SSB repeater:* AGC at 12-15 dB above noise floor; *Multi i-f LT:* AGC at 18-25 dB above noise floor. Certain i-fs, especially narrower, single-station ones, can have a 1-3 dB higher AGC setting. *Shared channel:* AGC at 25-30 dB above noise floor.

Measurements are made using 2.4-kHz instrumentation or receiver bandwidth. Noise output is directly related to the noise bandwidth by this formula: $dB\ change = 10 \log B_1/B_2$. Consequently, if a wideband output meter is placed on an LT with a 30 dB SNR in a 2.4-kHz bandwidth, the meter would see only a 20 dB SNR (if the LT's entire passband is 24 kHz wide).

i-f amplifier gain. The i-f amplifier must provide enough gain to bring the equivalent noise input of the preamp/converter up to a level that is 30 dB below the desired maximum output level (AGC maximum level). This assumes, of course, that our design goal is 30 dB AGC to noise output ratio. We've previously shown that the front-end gain consists of 25-30 dB preamp gain with a mixing loss of about 7 dB, leaving a front-end (down converter) gain of about 18-23 dB. For our purposes, we will assume 20 dB.

An additional factor must be considered. All preamps internally generate their own noise. Noise figure is the ratio of a preamp's actual noise output relative to what its output should be if it did not add any noise of its own to the system. In other words, while a perfect 10-dB gain preamp would give $-140\ dBm + 10\ dB$ or $-130\ dBm$ *noise output*, a typical low noise figure preamp would actually give about 2 dB more noise output, that is, −128 dBm. Consequently, a signal arriving only 10 dB above Johnson-Kelvin noise yields a 10 dB SNR with our perfect preamp, but only 8 dB SNR in our practical preamp.

The formula for determining the Equivalent Noise Input (N_{equiv}) of any receiver is:

$$N_{equiv} = -174\ dBm + 10 \log B_{Hz} + NF_{dB}$$

For our LT system, assuming a system noise figure of 3 dB and a bandwidth of 2400 Hz, N_{equiv} is:

$$N_{equiv} = -174 + 34 + 3 = -137\ dBm$$

For a noise output 30 dB below full output, our output noise level is:

$$N_{out} = P_{out_{max}} - SNR_{Desired}$$

Since we've settled on 30 dB for this case, and we've determined that 0 dBm is our $P_{out_{max}}$ i-f noise output is:

$$N_{out} = 0\ dBm - 30\ dB = -30\ dBm$$

The required i-f gain to achieve this noise output is:

$$G_{i\text{-}f} = N_{out} - (N_{equiv} + G_{DnCv})$$

For our example:

$$\begin{aligned} G_{i\text{-}f} &= -30\ dBm - (-137\ dBm + 20\ dB) \\ &= -30\ dBm + 117\ dBm \\ &= 87\ dB \end{aligned}$$

too much gain

Just as we've previously indicated that 60 dB or more gain in the *PA* stages begins to get difficult to

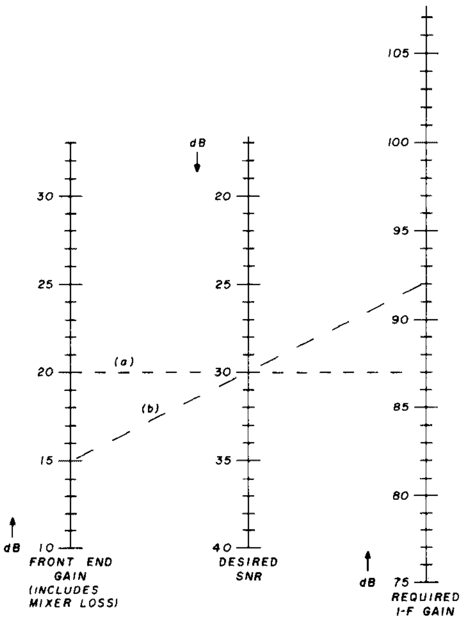


fig. 8. Nomograph illustrating the relationships between front-end, i-f gain requirements and desired signal to noise ratio.

control, gain exceeding 75-85 dB also begins to get unmanageable at most i-fs. This is why most fm radios use two i-fs, 10.7 MHz and 455 kHz, before detection. The amount of gain required to provide good fm limiting (about 90 - 100 dB at i-fs), exceeds the gain practical on any one given frequency. The same technique is applied to the LT design, that of splitting up the gain-producing stages at two different frequencies.

With a practical limit of 85 dB on any given frequency, a translator that uses a single i-f can set the noise floor no higher than about −30 dBm, or about 30 dB below full output. This gain is reasonable for multi-i-f or broadband single-i-f systems (assuming care is taken in shielding and layout), but cannot be pushed further for a repeater-type system (**fig. 8**).

a different approach

An approach I've wanted to try but have not yet had the opportunity to get to is the use of two lower gain i-fs on frequencies separated by the TX/RX offset split. The signal would be converted to 10.7 MHz, amplified by a factor of 50 dB, converted to 11.3 MHz (600 kHz up), amplified another 50 dB, and then converted to the output frequency.

This has two advantages. The same local oscillator can be used for both receive and transmit conversion, since the transmit i-f is already 600 kHz above the receive i-f. Also, improved frequency stability can be achieved using a subharmonic output from a stable oscillator. This means that any receiver LO drift is tracked precisely by the transmit LO (which is the same in this case), and drift of the i-f to i-f converter oscillator is divided by the same ratio used to obtain the 600-kHz injection signal. Example:

$$\begin{aligned} 144.510\,MHz - 133.810\,MHz &= 10.700\,MHz \\ 10.700\,MHz + 0.600\,MHz &= 11.300\,MHz \\ 11.300\,MHz + 133.810\,MHz &= 145.110\,MHz \end{aligned}$$

The 600 kHz can be derived from a 4.8-MHz crystal (divide by 8). Its drift at 4.8 MHz will likely be less than 100 Hz over moderate temperature excursions, thus the 600 kHz will drift only 100/8 Hz, or 12.5 Hz.

clamping the offset to 600 kHz

For in-band translators or SSB repeaters using normal offsets (600 kHz on 2 meters, 1.6 MHz on 220, and so forth), the one-frequency i-f scheme presents a slight problem in another area. If the offset i-fs are not used, the drift between the upconverter and downconverter LOs can be as high as 200-400 Hz. This is unacceptable on SSB.

NBC's solution was simple. The receiver's LO was made into a VCO (voltage-controlled oscillator) and slaved to the transmit LO after both were divided down to 600 kHz. Using this phase-locked loop (PLL) provides for nearly perfect 600-kHz offset with an error only as large as the drift of the 600-kHz reference oscillator divided by whatever divide ratio one is using. In fact, the NBC system uses a standard 10.240-MHz reference oscillator (as used in many CB sets and ham rigs), dividing everything down to a standard 10-kHz reference frequency using standard, well-proven CB-type PLL circuitry and parts. Thus our TX/RX offset accuracy is 10.240/0.600 MHz, or seventeen times better than that of the 10.240 MHz oscillator. Since this may drift less than 100 Hz under normal temperature conditions, our offset drift will typically run less than 100/17, or about 6 Hz (**fig. 9A,B**).

other problems

LTs have a new set of problems not generally seen on fm repeaters. The first problem is impulse noise. While an fm repeater in the presence of noise squelches more tightly, an LT will break squelch and retransmit that noise with the degree of fidelity its i-f selectivity allows. There are two methods of preventing this: 1. use a more sophisticated squelch system than the usual level detector, or 2. prevent noise from getting to the detector by using an effective form of noise blanking.

Single-channel systems (SSB repeaters, multiple-single channel systems) could use the ratio squelch systems developed by Kahn some years ago (**fig.**

10A and **B**). This device is sensitive to voice characteristics and triggers a squelch detector driven from an audio (or i-f) ratio detector or discriminator.

Single-channel systems can also use sub-audible carriers for COR (**fig. 10C**) or squelch triggering. If each station transmits a carrier, say, within ± 100 Hz of the frequency where the COR or squelch is tuned (whether an audio detector or narrowband 300-Hz i-f filter is used really doesn't matter), only signals at that frequency trigger the system. Noise, being broadband, needs to be much stronger to trigger the system.

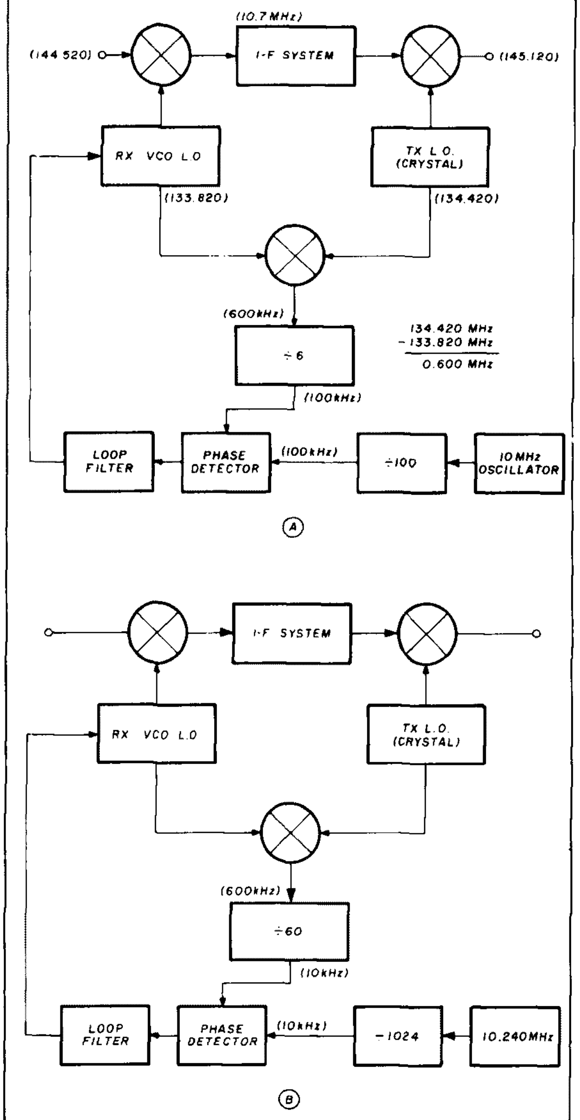


fig. 9. (A) NBC linear translator's solution to the drift problem uses the receiver's LO, a VCO slaved to the transmit LO through a phase-locked-loop. (B) NBC linear translator using an alternate PLL system with a 10.240 MHz reference oscillator.

If the entire passband of the LT is converted to audio frequencies, a series of af filters driving individual detectors can be used to cover the entire passband in 1-kHz steps (**fig. 10D**). If any one detector is triggered by audio in its frequency range, it will turn on the LT. The 1-kHz selectivity still provides some discrimination against triggering on noise by reducing it by 4 dB. Additionally, this provides greater sensitivity to weaker desired signals in the smaller bandwidths.

Another system uses two detectors with a comparator so that one averages noise, pulses, or signals present and the other sees the peak value (**fig. 10E**). CW, SSB, or fm have a fairly low peak to average ratio, so that if the comparator's gate signals were set to trigger only on signals whose peak to average value was less than 4:1, most pulse-type noise would not trigger it, since its peak to average is typically much higher than 4:1.

None of the above systems have been tried on the NBC LT. Reports on your experiments would be appreciated. We have chosen merely to monitor the system and reduce i-f gain when noise becomes a problem. This isn't elegant, but it works!

Noise blankers present a particular problem in LT service. While it is not difficult to produce a noise gate and the associated detector scheme, it is difficult to operate any NB system in a full-duplex system. The transmitter signal is usually strong enough, even after a duplexer, to cross-modulate the noise blanker.

Here are a few suggestions for resolving the noise blanker problem. The first is to detect and trigger on out-of-band noise pulses. Collins has used this technique for years and many CB sets have NBs tuned at 24 MHz rather than 27 MHz so that stronger signals aren't detected in the pulse detector circuit, causing crossmodulation products in the system. For 2 meters or other VHF/UHF frequencies one can tune just below or just above the band to find a 100-200 kHz range that is free of strong signals, convert it to an appropriate i-f, then build the pulse detector around that. A notch filter at the output frequency of the LT might also be required to protect the NB downconverter.

A more standard i-f-type noise blanker immediately following the downconverter of the LT could be used if run through a 30-kHz crystal filter to remove the output and strong adjacent signals. This will not be quite as effective due to the narrow bandwidth (pulses will tend to ring and broaden), but is used effectively in many 2-meter SSB rigs. Strong signals on immediately adjacent channels (± 20-30 kHz) may also cause some problems. Use of a 40-60 kHz filter

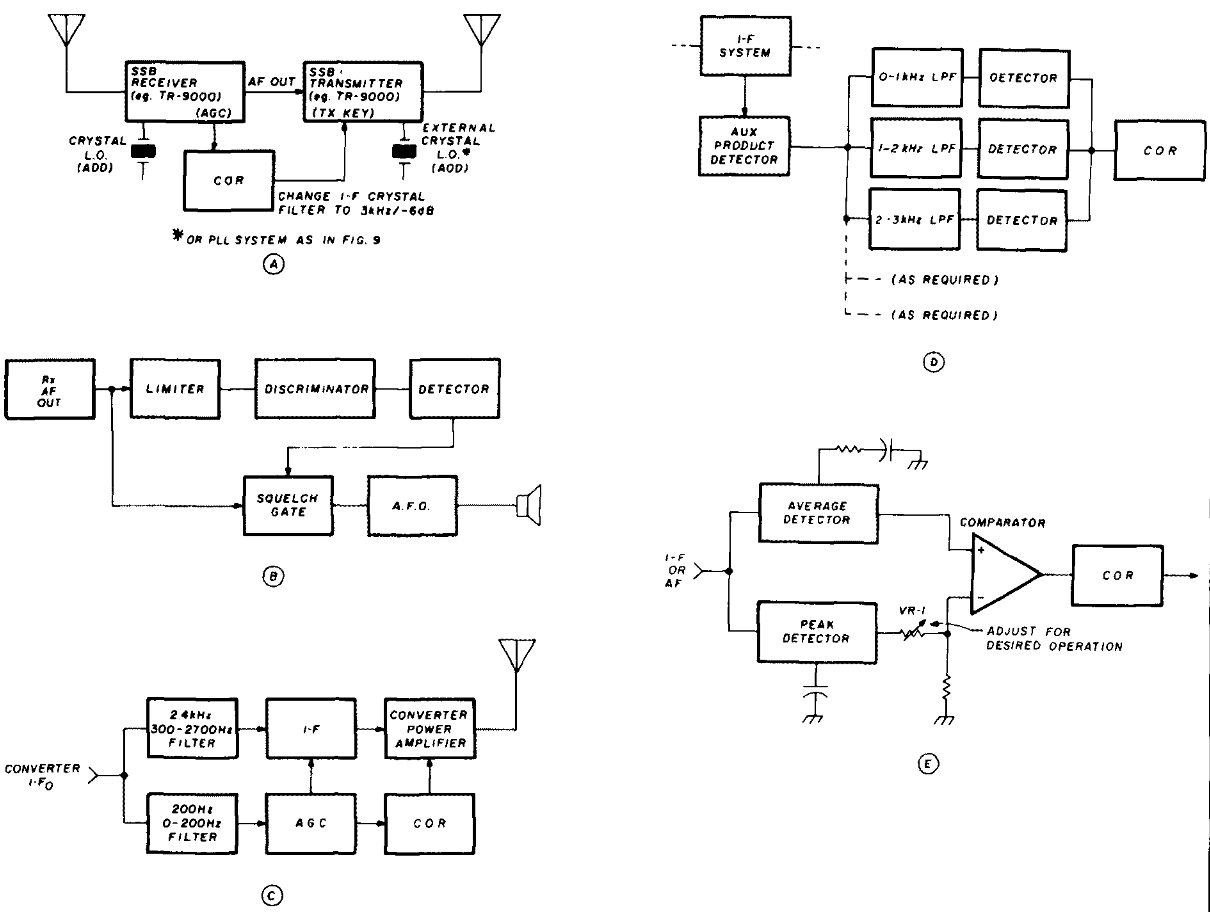


fig. 10. (A) SSB repeater block diagram. Use external TX and RX LO's or PLL system of fig. 9(A) and (B); (B) Ratio squelch design developed by Kahn is useful in both single and multi-channel systems; (C) Single-channel system use of sub-audible carriers for COR or squelch triggering. AF filtering off a detector or appropriate i-f filters can be used; (D) An alternate method of turning on an LT uses a series of 1 kHz bandwidth audio filters that drive individual detectors; and (E) An alternate COR system uses a comparator fed by an average and peak detector circuit. Decision is based on the ratio of average to peak levels received at the comparator.

might improve pulse sensitivity giving better blanking, but would increase the risk of adjacent channel crossmodulation.

With either technique some further crossmodulation protection can be obtained by using an averaging AGC having a slow attack and release characteristic. This will clamp CW, a-m, or fm signals (including SSB) when they start approaching the crossmodulation level, but will not clamp fully on quicker noise pulses. This technique is used quite effectively on many h-f ham rigs, but can be difficult to set up properly (as can be seen by other hf rigs having similar noise blankers that don't work). Fortunately, we are working with a single band and only a few kinds of noise on VHF, so that level setting of the AGC and its attack/release characteristic is not as critical.

I should point out that the reduction of gain which prevents crossmodulation of the noise blanker by signals will also reduce sensitivity to noise pulses so that such signals will at some point, in essence, turn off the blanker.

For the NBC translator we use an attenuator in front of the i-f amplifier to reduce the overall gain (prior to AGC) so that the normal noise floor and external interference are both dropped below COR threshold. This also reduces weaker station levels, but without some kind of noise blanker those stations can't be copied anyway. The argument in favor of the use of a noise blanker in spite of these technical problems is that they extend the weak signal capability of the LT by 6-12 dB in typical environments. Mountain-top locations may be quiet enough, however, to obviate the need for a blanker.

A comment on i-f clean-up filters should be inter-

jected here. Unlike repeaters whose bandwidth is automatically limited by de-emphasis and modulator bandwidth, an LT passes all i-f generated noise to the output frequency. Since the crystal filter is normally placed at the very front of the i-f strip to provide best IMD/selectivity performance, it does not limit the output of noise from any of the following i-f stages. The only selectivity limiting noise output from the i-f stages is provided by the interstage transformers and tuned circuits. At 10.7 MHz we see a noise bandwidth of about 120-250 kHz at the 3 dB points and only about 30 dB down at the 500-1000 kHz points. Needless to say, this is unacceptable at the output of the LT.

Fig. 11 illustrates several ways to improve this situation. The best (and most expensive) is to use an input and output filter on each i-f so that the noise output is essentially limited to the bandwidth of each i-f. A good compromise, though not best in terms of adjacent-channel and front-end and IM considerations, is to use a single filter as wide as all i-fs combined in front of the first couple of i-f stages, then split (or convert) the signal to individual i-f amplifiers each having individual filters. This controls the channel to channel IMD and noise outputs. Only noise from active channels can reach the upconverter and PA stages. Additionally, IMD products created in the first i-f will not spill over into the second i-f, and the output clean-up filter of i-f number 1 will not allow IMD products outside of its passband to be passed to the upconverter.

In areas of high signal congestion, better selectivity at the input may be more desirable. In this case, filters are placed at the input of each i-f with one clean-up filter at the output-combining point of the system. This causes broader i-f noise and IMD output and loses some of the advantages of shutting off unused i-fs to reduce output bandwidth.

Last, unlike fm repeaters, an LT will likely see positive feedback or regeneration when there is insufficient TX/RX isolation because of poor shielding, filtering, or inadequate duplexers. This shows up as what most people would term distortion, though it is really regeneration or noise pumping. Oscillations can also occur. Squelch and COR setting can also be aggravated as can the ratio of AGC to noise floor.

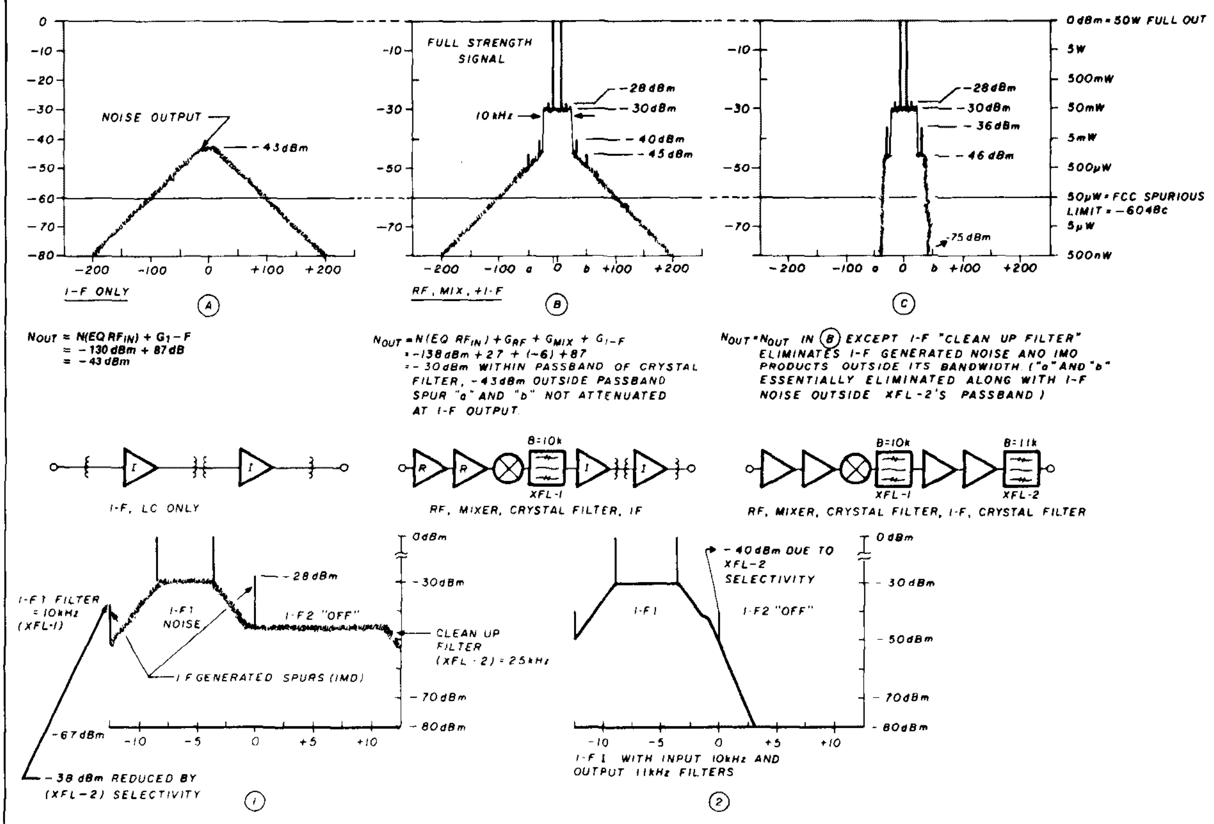


fig. 11. Filter techniques used to reduce the noise output of an LT.

summary

Current exploration of narrowband modes by the FCC, coupled with our understanding of the bandwidth, DX, and multipath advantages of SSB, now make the time right for hams, as the most experienced users of VHF/UHF SSB, to contribute once more to experimentation and expansion of the state-of-the-art. Building an LT is, at least, challenging, and, at most, could help develop a new technology.

I would be happy to assist anyone wishing to build an LT. For a complete set of schematics of the NBC LT, send $1.50 and a large SASE directly to me. I'll also respond to questions; just enclose an SASE with all inquiries.

notes

(a) FCC practice in fm and TV broadcasting is to give the name "translators" to stations that receive signals and change them to another frequency to extend coverage. AMSAT and many satellite agencies tend to use the term "transponder." My feeling is that after many years of such use in aviation, "transponder" has come to mean a device that returns a specifically coded signal when interrogated by a radar pulse; a "translator" merely translates, with no other modification or detection, an incoming signal from one frequency to another.

(b) The two meter, in-band Linear Translator is located in Twaine Harte, California and provides 50 watts ERP at 144.52/145.12 MHz in keeping with *suggested ARRL Two Meter Band Planning*. Information can be obtained by writing Neil Lewis, WB6VIV, 8119 Phaeton Drive, Oakland, California 94605. For specific technical information, contact the author. (SASE requested in either case.)

(c) This machine was taken off the air in 1981 for modifications and has been overhauled to act as a development LT for Project OSCAR's SYNCART satellite project. It should be back on the air by the time this is published, with input at 1296.3 MHz and 145.7 MHz with output at 435.415 MHz. Inquiries concerning it should be addressed to Project OSCAR, c/o the author at his home address. (SASE required.)

(d) Fm receiver selectivity = 18 kHz, typical SSB receiver selectivity = 2.4 kHz, typical SNR improvement (SSB over fm) = 10 log 18/2.4 = 9 dB.

(e) Note the 9-dB bandwidth advantage (particularly below threshold for fm) coupled with the additional 6 to 18 dB SNR improvement given by use of ACSB. This offsets most, if not all, of fm's noise-limiting capability.

(f) Sensitivity improvement = 10 log 20/10 = 3 dB.

(g) This proposal has been reworked recently by AMSAT Canada and Project OSCAR. The original host launch was cancelled due to delays in the Space Shuttle program. Two possible upcoming launches will not allow time for a SYNCART-style transponder to be developed. Work continues, however, on SYNCART development for possible later launches.

(h) This section, although describing an LT design, can be used to improve fm repeater designs.

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ham radio

a fail-safe amplifier power supply

An improved method of providing dc voltages to GaAs FET amplifiers

It has long been recognized that certain semiconductor devices require a pair of voltages of opposite polarity for proper operation. Amateurs have begun using one of these devices, the gallium-arsenide field-effect transistor (GaAs FET), in low-noise preamplifiers on the bands above 100 MHz. But the (N-channel) GaAs FET requires that a negative voltage always be present between its gate and source leads whenever a positive voltage is present between the drain and source leads of the same device. Failure to provide the negative gate-source voltage ($-V_{gs}$) before the positive drain voltage ($+V_{ds}$) is provided leads to the destruction of the device.

single dc lead

The problem of providing gate-source bias voltage at a time no later than the drain-source voltage is provided, is rather easily solved if both positive and negative voltages are available for the amplifier. However, because full advantage of the low noise-figure of a GaAs FET preamplifier is obtained only if the amplifier is mounted at the antenna feed, it is often undesirable to provide three dc wires (+ voltage, − voltage, and ground) to the antenna-mounted preamplifier. A preamplifier can be operated with a single positive voltage if some form of self-biasing configuration is used. Typically, self-bias is provided by a resistor from the device source lead to ground; however, a very low-loss bypass capacitor must parallel the source resistor to effectively place the device source lead at rf ground potential. The bypass capacitor is generally of the chip capacitor type, and may often be self-resonant at the Amateur band of interest.[1] However, as the frequency of use is increased to beyond about 3 GHz, even the best available bypass capacitors add undesired reactance; the GaAs FET device is best operated with the source lead connected directly to the circuit ground. When this connection is made, the pair of opposite-polarity voltages are again required, and the problems of power supply sequencing and number of supply wires are again encountered. Because it is very desirable to provide a single voltage, generally of positive polari-

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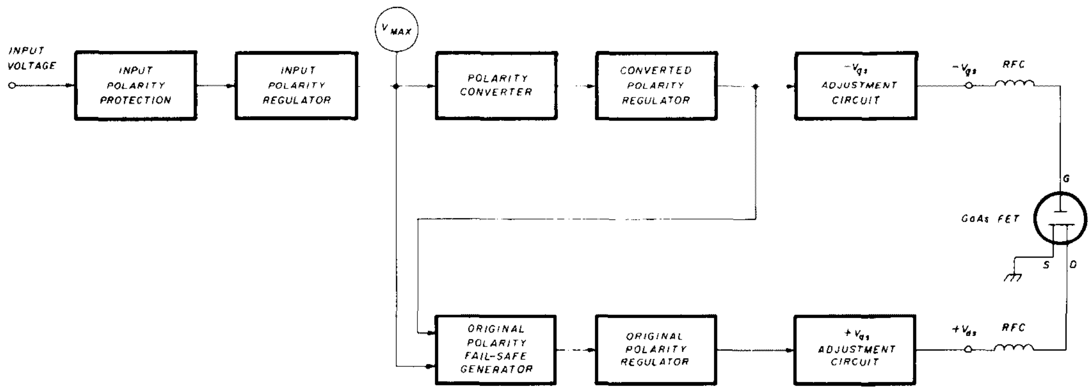


fig. 1. Block diagram of a fail-safe power supply for a GaAs FET low-noise preamplifier.

ty, along the inner conductor of the coaxial cable to the preamplifier output, with the coaxial cable outer shield providing its normal ground connection, some form of power supply at the preamplifier is necessary to provide both polarities of voltage in proper sequence.

My fail-safe power supply (**fig. 1**) was developed to solve this problem (and was subsequently found to be novel enough to be granted a patent in the United States). It is, in some ways, very much similar to the power supply described by Norman Foot, WA9HUV, in an earlier issue of this magazine.[2] My patented circuit has several features worth consideration by anyone planning to power a GaAs FET preamplifier from a single polarity voltage supply.

GaAs requirements

Consider first the voltage and current requirements of a typical (low-noise) GaAs FET device: V_{gs} usually must never exceed -6 volts, at a very small current (generally less than 1 microampere); V_{ds} is generally about $+3$ volts at about 10 milliamperes in operation and must generally never exceed $+6$ volts or a maximum current of 50-100 milliamperes. It is important to reduce the supply voltage to a voltage no greater than the 6-volt maximum voltage, as soon as possible in any proposed power supply. This maximum voltage V_{max} is provided by an input polarity regulator, utilizing a 78L05 integrated circuit. The input polarity regulator is itself protected from receiving an input voltage of improper polarity by the use of a so-called "idiot diode", CR_1. An input capacitor, C_1, is used to prevent input voltage surges when the power supply is in the presence of a larger rf field (as may occur with the power supply at the antenna). The input polarity regulator device U_1 is suitably bypassed and is protected from reverse voltages, during power-down, by diode CR_2. The net result is a positive voltage at test point TP1 which is less than the maximum safe drain voltage, provided at a very early point in the power supply, and protected against both reverse-polarity and overvoltage faults by diodes CR_1 and CR_2.

The positive voltage is converted to a negative voltage in a polarity converter using a relatively new integrated circuit, U_2, the ICL7660 from Intersil, Inc. The ICL7660 requires only two associated components, the pair of 10 μF capacitors C_4 and C_5, in order to generate a negative voltage at test point TP2 of about -4.5 volts. While I have tried the polarity converter circuit using a 555 timer integrated circuit, as suggested by WA9HUV (and the 555 timer circuit does work as intended), the three-component circuit of **fig. 2** is somewhat simpler, more compact, and more reliable. The negative output voltage is of almost the same magnitude as the positive input voltage and requires neither a voltage doubler nor operating voltages greater than the maximum voltage of the GaAs FET, a preferred feature in case of component failure in the polarity converter.

Because the amplifier noise figure and, to a somewhat lesser extent, the amplifier gain are directly related to the device drain current I_d (set by the gate voltage V_{gs}), the converted-polarity voltage must be highly regulated. The converted polarity regulator uses a series resistor, R_1, and a zener diode, CR_4, to provide this regulated voltage; -3.3 Vdc is provided at test point TP3. The zener diode CR_4 also prevents a positive voltage from ever appearing at the device gate. CR_4 is shunted by filter capacitor C_6 to remove any noise components generated by the zener. The regulated negative voltage appears across potenti-

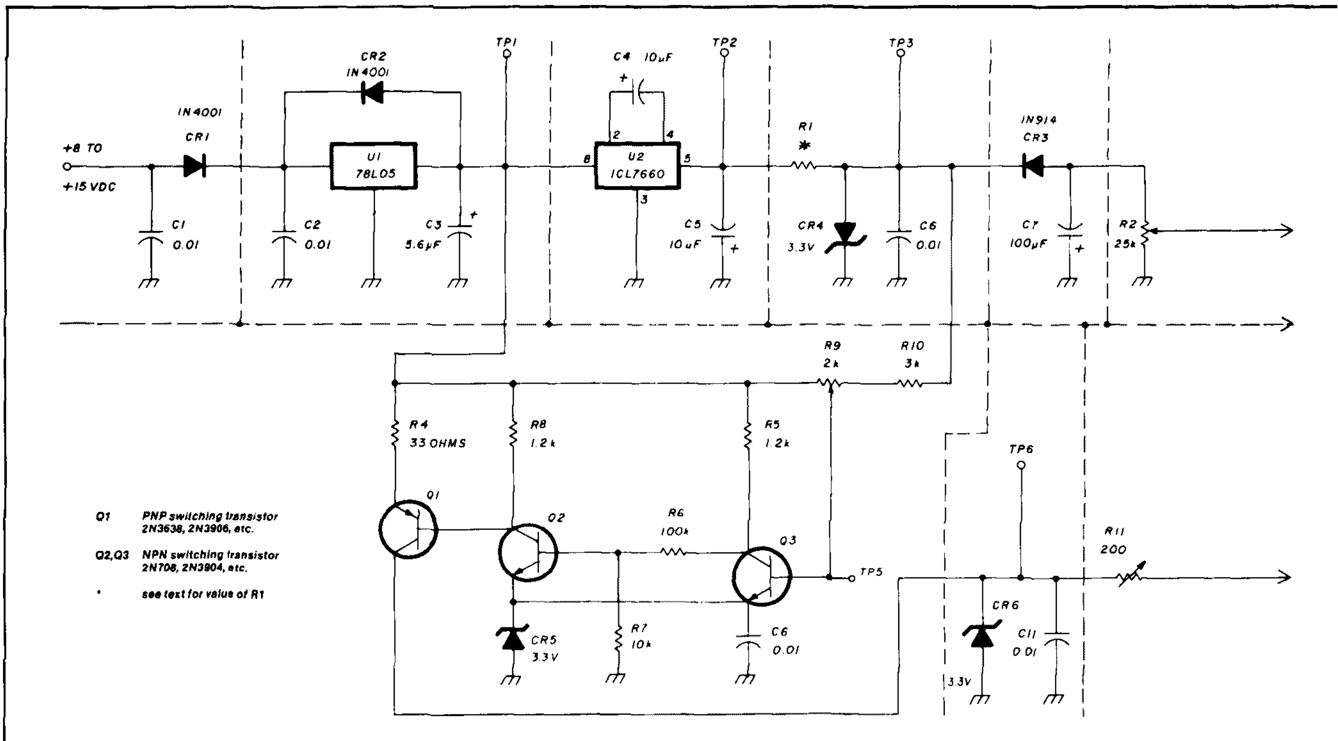


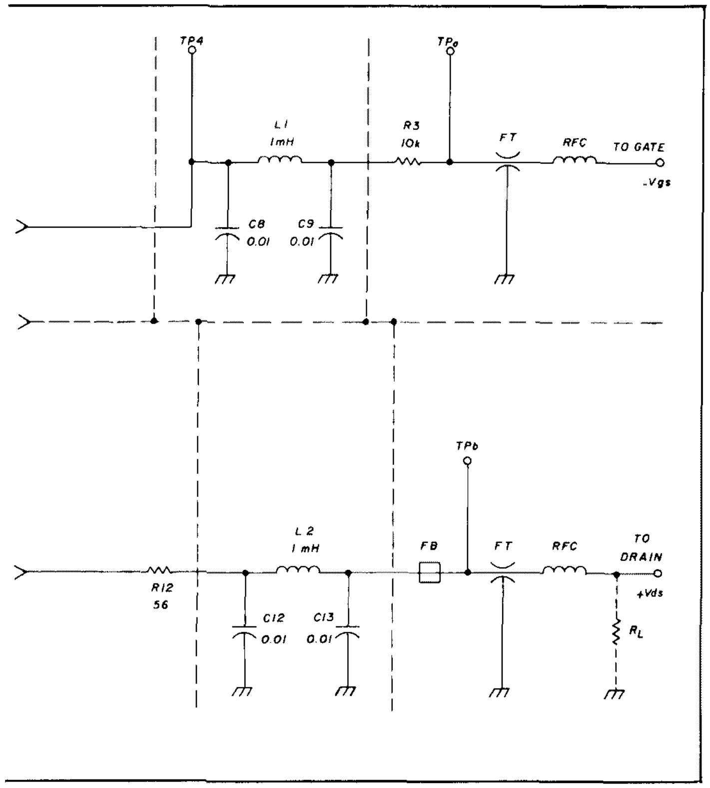
fig. 2. Circuit diagram of a fail-safe power supply for a GaAs FET low-noise preamplifier.

ometer R_2, which is adjusted to set the amplifier device voltage $-V_{gs}$.

failure mode protection

The input-regulated voltage V_{max} and the converted-polarity regulated voltage at test point TP3 are used in an original-polarity fail-safe voltage generator subcircuit to provide a positive voltage for the amplifier device drain circuit *only* if the negative amplifier device gate voltage is present. Reason dictates that a switch, transistor Q_1, is open until the negative voltage at TP3 is present; that is, when the Q_1-Q_3 circuit of **fig. 2** fails in an open condition, with voltage and current not supplied to the subsequent original-polarity regulator subcircuit, which uses zener diode CR_6. Compare this approach to that of the crowbar/regulator U_2 circuit discussed in WA9HUV's article, in which failure of the crowbar circuit no longer reduces positive regulator U_2 input and thereby allows positive drain voltage to continue to be supplied to the amplifier device. My original-polarity fail-safe generator uses a Schmitt-trigger circuit (transistors Q_2 and Q_3, zener diode CR_5 and resistors R_5-R_8). The switching input voltage of the trigger is determined by the setting of potentiometer R_9, and operates as follows: if the proper -3.3 Vdc is present at TP3, then the voltage at trigger input test point TP5 can be adjusted to be sufficiently close to the zener voltage of diode CR_5, so that transistor Q_3 is cut off. The voltage at the base of transistor Q_2 is then greater than the voltage across zener diode CR_5, and transistor Q_2 is saturated, turning on transistor Q_1 and allowing a current, set by resistor R_4, to flow into zener diode CR_6. This provides a regulated $+3.3$ volts at positive regulator test point TP6. If the negative voltage at test point TP3 decreases, for any reason, by an amount determined by the setting of potentiometer R_9, the voltage at test point TP5 increases (becomes more positive) by an amount adjusted to raise the base of transistor Q_3 above the voltage of zener diode CR_5. Transistor Q_3 switches into saturation and transistor Q_2 switches into the cut-off condition, cutting off current flow through transistor Q_1 and reducing the voltage at positive regulator test point TP6 toward zero.

Resistor R_1 can be selected, in the 100-500 ohm range, to cause one of two results to occur when the fault condition causing the negative regulator test point TP3 voltage to be less than the design value is cleared. For lower values of resistor R_1, clearing of a fault (such as a short across zener diode CR_4) causes the trigger to reset and the circuit to return to normal



operating condition. For higher values of resistor R_1, clearing of a fault and return of the TP3 voltage to about -3.3 Vdc can cause the positive drain voltage to remain at about 0 volts until the main input power is removed and reapplied. This latter condition, somewhat dependent upon the setting of potentiometer R_9, is observable if the total current into the power supply is monitored. It can be useful in providing an automatic fault alarm, even while the device is being protected by the application of a safe negative V_{gs} and simultaneous removal of V_{ds}.

The amplifier device drain voltage V_{ds} is set by potentiometer R_{11} in conjunction with fixed resistor R_{12}. Typically, the drain voltage V_{ds} has an effect upon the amplifier noise figure and gain, although this effect is somewhat less than the effect of varying the device drain current, which is varied by adjustment of the gate voltage $-V_{gs}$ at potentiometer R_2.

One problem which will be encountered, but was not addressed in WA9HUV's article, is the possibility of voltage spikes occurring at the oscillation frequency of the polarity converter subcircuit. These spikes are generated by the squarewave switching inherent in either the 555 timer or ICL7660 converter integrated circuits. The first step necessary to prevent these spikes from showing up at either the amplifier $-V_{gs}$ input at test point TPa, or the drain voltage $+V_{ds}$ input test point TPb, is to build the power supply in a separate shielded box mounted next to the preamplifier shielded box. The negative and positive voltages are then run through short, shielded wires to the preamplifier input feed-through capacitors FT. As a second step, the converter "hash" is suppressed by a π-section lowpass filter (L_1, C_8 and C_9 or L_2, C_{12} and C_{13}) placed after adjustment potentiometers R_2 and R_{11}. Ideally, each of the filters should be in its own little shielded enclosure within the main power supply shield enclosure.

The additional circuit of diode CR_3 and capacitor C_7 is an added protective measure. During normal operation the voltage drop across diode CR_3 will provide a maximum gate voltage of slightly less negative magnitude than the zener voltage of diode CR_4. The voltage drop across diode CR_3 is substantially constant and the initial setting adjustment of potentiometer R_2 takes this voltage drop into account. However, during power supply turn-off, under either normal operation or because of a decrease in the voltage at TP3, the previously-charged storage capacitor, C_7, maintains a sufficiently negative voltage at the gate output to prevent destruction of the protected device while the voltage at the drain connection rapidly falls to zero. Typically, drain voltage will fall to zero in under ten milliseconds, while the voltage at gate test point TP4 requires about one second to fall to zero magnitude, providing an additional degree of GaAs FET protection.

circuit test

The circuit test should be carried out before the power supply is connected to the companion preamplifier. A dc load resistor, R_L, is connected to approximate the desired amplifier device drain load. For a typical 3.0 Vdc drain voltage at a typical 10 milliampere drain current. R_L is a 300 ohm, 1/8-watt resistor. Since the amplifier device gate electrode does not draw any appreciable current, the gate end of series-protection resistor R_3 is left open.

Apply a positive voltage in the 8 to 15-volt dc range to the power supply input and check for the following voltages, using a voltmeter with at least 20K ohm resistance (all voltages are ± 10 percent): $+5.0$ Vdc at TP1; about -4.5 Vdc at TP2; and -3.3 Vdc at TP3. The voltages at TP4 and TPa should be variable from zero to about -2.8 Vdc, which is less than the voltage at TP3 because of the drop in diode CR_3.

The positive-polarity portion of the power supply is tested by adjusting potentiometer R_9 until a 3.3-volt dc level appears at TP6. The operation of the fail-safe generator subcircuit is tested by temporarily placing a short circuit from TP3 to ground and noting that

the voltage at TP6 falls to zero. Some adjustment of potentiometer R_9, while monitoring the TP5 voltage, may be required to obtain the proper operation of the trigger circuit. Potentiometer R_{11} is then adjusted for the desired target drain voltage at TPb, with the dummy drain resistance, R_L, connected.

As a final check before connecting the power supply to the amplifier, monitor the TPa and TPb voltages with an oscilloscope (preferably starting at one volt per division and working down to the greatest sensitivity possible) and look for spikes in the kHz repetition frequency range. If such spikes should be found, increased values of the lowpass filter components may be necessary.

applications

The fail-safe preamplifier power supply has been used with, among other circuits, a π-network input/output preamplifier for 1296 MHz.[3] A number of different GaAs FET devices were tested; optimum noise performance could be obtained by varying the potentiometers R_2 and R_{11} in a noise-figure measurement test setup. The same devices in the same preamplifiers were also tested with a source-resistance-biasing scheme; a small but discernable increase in noise figure was found above the case where the device source leads were directly grounded and the fail-safe power supply used.

A very-low-noise receiver was built at 3456 MHz, using a simple single-balanced mixer[4] built on G-10 printed circuit board stock and garden-variety mixer diodes (HP2810). While the use of G-10 board is not recommended at this frequency, and resulted in a mixer conversion loss of about 13 dB (and a noise figure estimated at about 15 dB), a two-stage GaAs FET preamplifier with 29 dB of gain completely overcame the mixer noise. The preamplifier is almost identical to a well-known TVRO preamplifier,[5] with the exception of an NE21889 device used in the input stage and improved input matching through the use of a pair of 1/8 inch wide by 1/4 inch long copper foil "flapper" capacitors. The power supply recommended by the manufacturer requires a +15 volt supply and a −15 volt supply, in addition to a pair of 747 dual operational amplifier integrated circuits, a handful of zener diodes, and other components. In my receiver, that power supply was replaced by a pair of power supplies as shown in **fig.2**, allowing not only the independent setting of the drain voltages of each stage (which could not be done with the manufacturer's recommended power supply), but also the independent setting of the drain current of both stages, to provide a preamplifier noise figure of about 1.0 dB at the indicated 29 dB gain figure. Application of the usual series-stage gain and noise figure formulas will show that the overall converter (preamplifier plus mixer) gain is about 16 dB, with a total noise figure of about 1.1 dB. This illustrates that the poor mixer specs are, in fact, overcome by the superior gain of the preamplifier.

conclusion

A power supply circuit requiring only a single polarity of input voltage, yet offering a high degree of fail-safe protection for GaAs FET amplifiers — especially for the microwave Amateur bands — is now available. Any questions addressed to the author accompanied by an SASE will be answered.

license grant

United States patent laws prevent anyone, unless specifically licensed by the patent owner, from "making, using or selling" any circuit covered by the claims of my U.S. patent 4,320,447; this includes the circuit of ***fig. 2*** *as well as the circuit used in reference 2. The patent owner, being the author of this article, therefore expressedly grants a non-assignable license to any licensed Amateur Radio operator in the United States of America, to make and use the fail-safe amplifier power supply covered by the claims of U.S. Letters Patent 4,320,447 for use with amplifiers operating only in the authorized U.S. Amateur Radio bands and only for Amateur Radio activity. This limited license cannot be sub-licensed. Use of the circuit covered by the claims of Patent 4,320,447 with amplifiers operating on non-Amateur bands or for non-Amateur use, or the sale of the circuit for any use, are all expressedly excluded from this license; grants and terms of such licenses can be discussed directly with the patent owner.*

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2. Norman Foot, WA9HUV, "Safe Power for Your Low-Noise GaAs FET Amplifier," *ham radio,* November, 1982, page 18.
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4. D. Evans, G3RPE, and G. Jessup, G6JP, *VHF/UHF Manual, 3rd Edition,* RSGB, 1976, page 4.41.
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ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

Just about the time you are reading this column, the summer static level should be dropping off and the low frequency DX bands should be coming back to life. Low frequency DX conditions should improve during the next few years as the sunspot count continues to drop. This means that the 160 meter band will be a good performer for a number of years. Some Amateurs have stuck to 160 meters, year in and year out, but many of the newer Amateurs are not yet aware of the DX possibilities of this band.

A majority of the new transceivers have incorporated the 160 meter band and the 1983-84 winter season promises to be a good one for this venerable Amateur band, with a high level of activity.

the 1983/84 DX season on 160 meters

The number one DX operator on 160 meters is Stew Perry, W1BB, who has been active in that portion of the spectrum since 1920! (See **figs. 1** and **2**.) I asked Stew what he thought about the forthcoming season and this is what he said:

fig. 1. Stew Perry, W1BB, on the alert for 160 meter DX in 1920. Transmitter is a single tube oscillator, modulated for a-m phone. Running 35 watts, Stew quickly discovered the thrill of working DX.

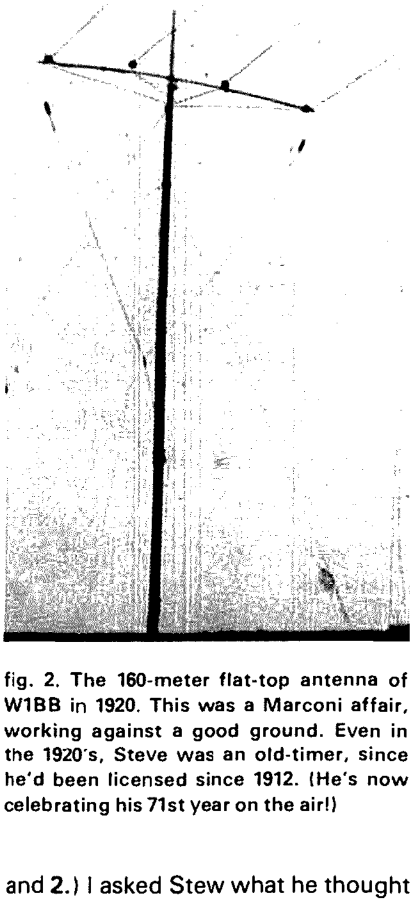

fig. 2. The 160-meter flat-top antenna of W1BB in 1920. This was a Marconi affair, working against a good ground. Even in the 1920's, Steve was an old-timer, since he'd been licensed since 1912. (He's now celebrating his 71st year on the air!)

> While I've not been very active recently, I've made a few good contacts this year (4X4NJ-Israel, PY1ARS-Brazil, and others). I think conditions seem quite sound and feel "right" for a good season. In spite of some changes, it still is the "gentleman's band", where folks are friendly and cooperative, without so much push and shove!
>
> By its nature, you're forced to work DX the hard way, at all hours of the day and night for world coverage. Sunset and sunrise (local time) are excellent periods, full of good DX surprises.
>
> As far as antennas for the "top band," the best receiving and transmitting antennas are required for top results, but you will be surprised what can be done with a simple inverted-L Marconi, or a Sloper with a good ground system.
>
> Come on down to 160 and meet some fine guys.

FCC rules and regulations

A collection of the FCC rules and regulations governing ham radio fills a thick notebook! Is it all worth the effort? Let's return again to the simple regulations of the Federal Radio Commission of 1927 (**fig.3**). Think of how many problems these simple rules would solve. As a famous philosopher once said, "I have seen the past, and it works!"

horizontal versus vertical antennas

I am still getting mail on my comments concerning horizontal versus vertical polarization for simple high frequency antennas (October, 1982, and April, 1983). A recent letter from my friend Win Wagener, W6VQD, says, in part:

> Too many people discuss the pros and cons of polarization without a good understanding of what is involved. It is not a simple question of which is better, a vertical or a horizontal antenna; but a need to consider the full environment of the antenna system.
>
> Too many articles on the effect of the surface of the earth in the vicinity of an antenna assume a perfectly conducting and flat surface for the

October 28, 1927.

Supervisors of Radio and Others Concerned:

For your information and guidance the Federal Radio Commission has established the following regulations governing the licensing and operation of amateur radio stations:

Amateur radio stations are authorized for communication only with similarly licensed stations and on wavelengths or frequencies within the following bands:

Kilocycles	Meters	Kilocycles	Meters
401,000 to 400,000	0.7477 to 0.7496	8,000 to 7 000	37.5 to 42.8
64,000 to 56,000	4.69 to 5.35	4,000 to 3,500	75.0 to 85.7
16,000 to 14,000	18.7 to 21.4	2,000 to 1,500	150.0 to 200.0

and at all times unless interference is caused with other radio services, in which event a silent period must be observed between the hours of 8:00 p. m. and 10:30 p. m., local time and on Sundays during local church services.

Amateur radio telephone operation will be permitted only in the following bands:

Kilocycles	Meters
64,000 to 56,000	4.69 to 5.35
14,500 to 14,000	20.68 to 21.4
2,000 to 1,580	150 to 190

Spark transmitters will not be authorized for amateur use.

Amateur stations must use circuits loosely coupled to the radiating system or devices that will produce equivalent effects to minimize key impacts, harmonics and plate supply modulations. Conductive coupling, even though loose, will NOT be permitted, but this restriction shall not apply against the employment of transmission line feeder systems to Hertzian antennas.

Amateur stations are not permitted to communicate with commercial or government stations unless authorized by the licensing authority except in an emergency or for testing purposes. This restriction does not apply to communication with small pleasure craft such as yachts and motor boats holding limited commercial station licenses which may have difficulty in establishing communication with commercial or government stations.

Amateur stations are not authorized to broadcast news, music, lectures, sermons or any other form of entertainment.

No person shall operate an amateur station except under and in accordance with an operator's license issued to him by the Secretary of Commerce.

W. D. TERRELL.

rk

Chief, Radio Division.

fig. 3. Early FCC regulations were simple, straightforward.

mathematical calculations. Unless the antenna is located above salt water (bay, tidal land, or ocean), which gives a flat, low-loss surface, the real earth is usually quite different and does not meet this requirement.

Calculations for the radiation intensity at different vertical angles from an antenna are based on combining the direct radiation from the antenna at a particular wave angle with the wave reflected from the ground. Wave reflection takes place at a distance from the antenna which varies with the particular angle of elevation being studied. For a high angle the reflection area is near the antenna, but for low angles the area of reflection is more remote, and for nearzero takeoff, that area is many wavelengths from the antenna.

For vertical polarization at low angles, the absorption of energy by the earth increases with ground loss for some distance from the antenna. There is considerably less ground loss upon reflection when antenna polarization is horizontal.

In addition to the need in real life to consider the effect of ground loss, the surface of the earth is often not flat, but hilly out to a thousand feet or so from an antenna. This must be taken into consideration. If, for instance, the radiation from an antenna system fifty feet above a flat surface is desired at an angle of five degrees elevation, the earth surface involved in the reflection area is five hundred to six hundred feet away from the antenna. If the ground slopes away from the antenna, the reflection area is closer and if the ground rises, the area is further away.

In my case, near the top of a ridge with thin topsoil and shale below, slopers and phased verticals seemed always below the performance of a halfwave horizontal antenna.

Win brings up a good point. Antenna reflection drawings in handbooks and articles are comforting, but they assume a perfectly conducting ground surface. In actuality, the "lay of the land" within five hundred feet, or more, of your antenna determines the actual reflection pattern.

Many Amateurs, surrounded by other people's houses, telephone and utility wires, and television antennas, can only guess at the angle of takeoff of their signal as the reflecting ground surface is obscured. So don't take

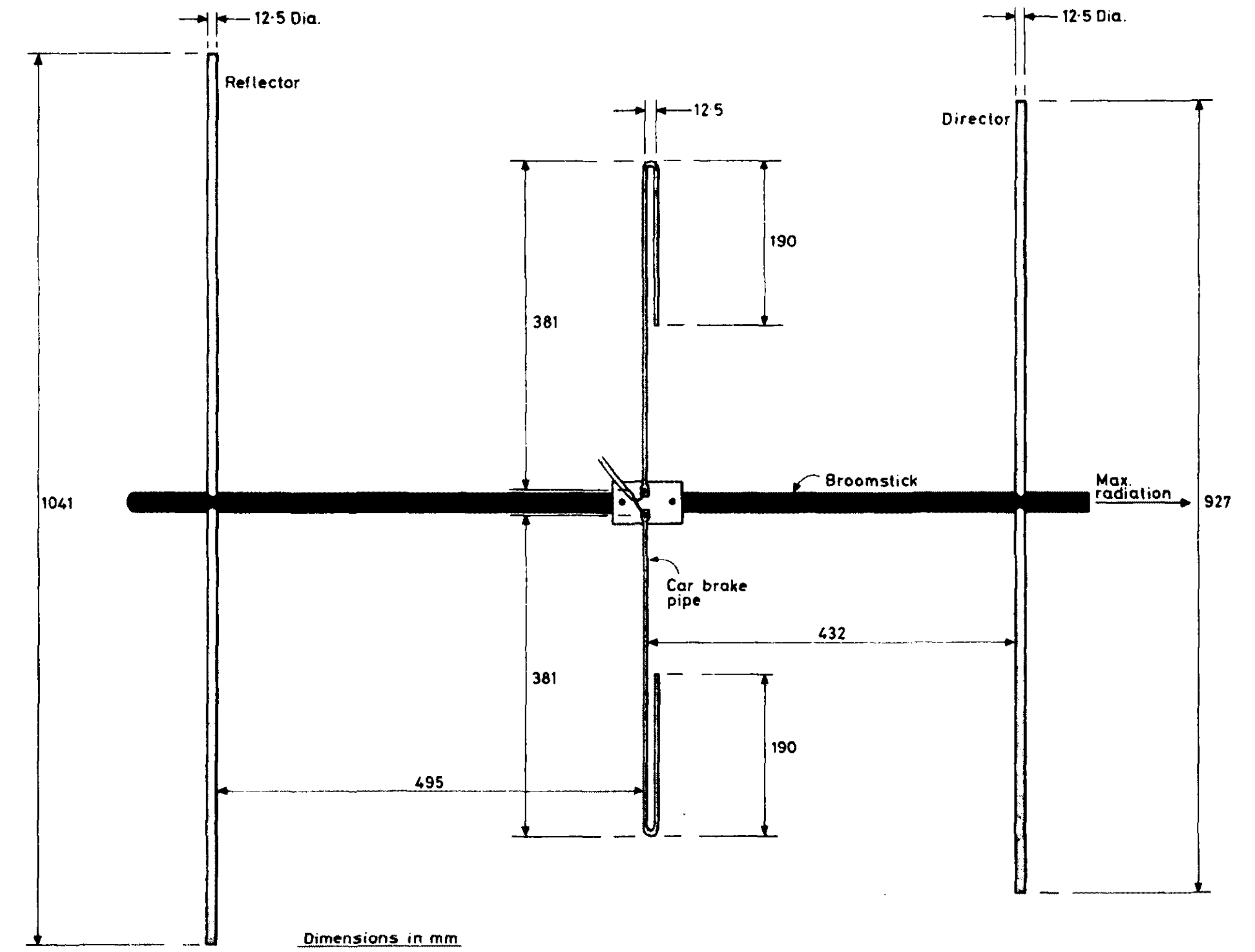

fig. 4. The unusual 2 meter Yagi beam designed by G6AFJ. Note that the driven element is folded back upon itself at the tips. Does this provide a good match for the 50 ohm transmission line? Try it and see! (Dimensions are in millimeters.) Illustration from *Practical Wireless*.

the patterns seriously. In most cases, the higher the antenna, the better the DX results. And that's probably due less to the reflection angle of takeoff than to elevating the antenna above nearby conducting structures!

the G6AFJ beam antenna for 2 meters

Is someone pulling my leg? Or does it really work? (See fig. **4**.) This is reproduced from the February issue of *Practical Wireless,* a well-known British publication. (Antenna dimensions are given in millimeters.) The noteworthy aspect of this three-element Yagi for 144 MHz is the unusual driven element, with its tips folded back upon themselves, presumably to provide a match to the 50-ohm line. G6AFJ says "the antenna can be tuned by altering the position of the bends in the folded elements to vary the gap."

The driven element is made of "car brake pipe," a substance that is unknown to me, but I would assume that it is thin-wall copper tubing. It appears to be about 3/16 of an inch (5 mm) in diameter.

In any event, the antenna is simple to make, and if any readers try out the idea, I'd like to hear about the results.

inexpensive station clock

Would you like a small, accurate station clock that you can set to WWV and buy for less than $15.00? I found one at my local hardware store. (I also found the same item for sale in several drug stores.) I am talking about the Timex model 5204-412 digital clock that sells for $9 to $14. This compact clock has a large red LED display of hours and minutes, plus an indicator of "a.m." and "p.m." It also has an alarm which is handy for keeping to schedules.

The instruction manual tells everything about the clock except how to set it to WWV, so the "minute" LED advances exactly on the proper WWV time-tick. Once you know how to do it (and I found out by experimentation), it is easy to lock the clock within a second of WWV. Here's the trick. Advancement of time as shown on the "minute" LED is performed when the clock is turned *on*. For some obscure reason (don't ask me why) there is a built-in 30 second time-shift in the clock. The trick is to plug the clock into your power receptacle exactly 30 seconds *after* a WWV minute tone. If you do this, then when you set the clock to WWV the LED will advance exactly on the minute tone of WWV. Once you accomplish this simple feat, the little clock runs right along with WWV UCT time.

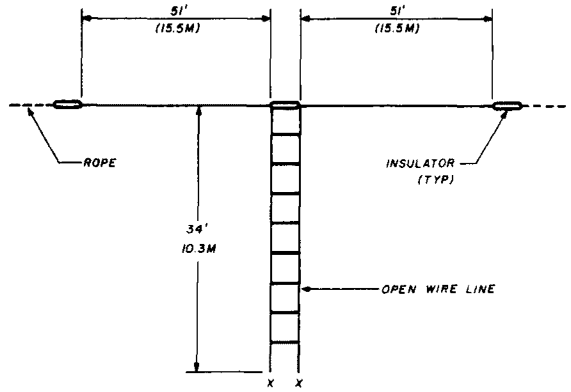

fig. 5. The G5RV multiband antenna. This simple device has been operated on all Amateur bands between 160 and 2 meters! One version of the antenna connects a 50 ohm coaxial line at points X-X and uses an antenna matching unit at the station to reduce the SWR on the line to a low value suitable for the transmitter. Another version extends the open wire line to the transmitter and uses a balanced antenna tuner to match to the transmitters coaxial output.

the G5RV "all band" antenna

Have you ever noticed the antenna description on an overseas QSL card was the "G5RV antenna?" Little-known in W-land, the G5RV is a popular multiband wire antenna used by many overseas operators. Popularized by Louis Varney, G5RV, the basic antenna is shown in fig. **5**.

A 102 foot long flat-top is used, fed at the center with a length of open wire transmission line. The antenna operates as a shortened dipole on 80 meters, an extended dipole on 40 meters, nearly three 1/2 wavelengths on 20 meters, and as a center-fed long wire on 15 and 10 meters.

There are many methods of feeding the G5RV antenna. The original design uses an open wire stub plus a length of coaxial line. An antenna tuning unit is used in the station, as the SWR on the coaxial line can be quite high at certain frequencies.

A second feed system is to extend the open wire line directly into the station, and use a balanced antenna tuner at this point.

The G5RV feeders can be connected in parallel at the bottom of the stub and the antenna operated against ground as a top-loaded vertical on 160 meters. Some hams have even used it on 6 and 2 meters by using a VHF antenna matching unit at the station!

My preference is to run open wire line from the antenna directly to an antenna tuner in the station. (Saxton makes heavy-duty open wire line, I believe.) For low power (a hundred watts or so), TV-type twin lead may be substituted for the open wire line.

moonbounce revisited

The popular brochure "All You Want to Know About Moonbounce (EME) Transmission" has been reprinted and is once again available. If you wish a copy, send four 20¢ stamps, or four IRCs, to me at the following address: Eimac Division of Varian, 301 Industrial Way, San Carlos, California 94070.

ham radio

rf synthesizers for hf communications: part 2

Understanding the phase-locked loop

Part one described the functional blocks of a PLL synthesizer and loop operation and presented a simple approach to PLL synthesizer design. This part details the loop components with an analysis that points out aspects critical to an understanding of the loop. The open loop performance is then examined in terms of its individual sections which leads to a comparison of open and closed loop operation. Finally, phase noise basics are discussed, showing the effects of a PLL on closed loop VCO phase noise.

VCO operation

VCO gain defined by $K_{VCO} = \Delta f_{VCO}/\Delta V_t$, indicates the dependency of VCO frequency on a tune or control voltage. However, a VCO when used in a phase locked loop incorporates a phase not a frequency detector. Therefore the effect of ΔV_t on VCO phase, relative to the reference oscillator phase (f_{REF}), is seen as a system integration.

Fig. 1A shows a perfectly stable reference oscillator at a frequency of f_{REF}, and a VCO that has a $K_{VCO} = 1\ MHz/volt$. The phase detector is a perfect phase detector with an input range of zero degrees to infinity and a $K_\phi = \frac{1\ volt}{10\ degrees}$. If f_{REF} and f_{VCO} equal exactly 5 MHz, and are in phase, then the phase detector output will be 0 volts.

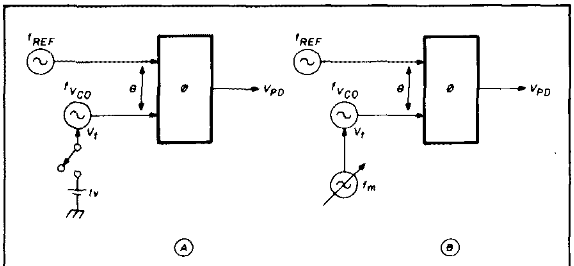

fig. 1. Circuit block arrangement to show combined VCO-phase detector response to a voltage step and swept frequency on V_t.

At $t = 0$ a plus 1 volt applied to the VCO tune line causes an instantaneous 1 MHz jump in VCO frequency. 100 nanoseconds after voltage is applied the phase difference between f_{REF} and f_{VCO} is:

$$\theta = 360° t(\Delta f)$$
$$\theta = 360° (100\ ns)(1\ MHz) = 36°$$

At 1 μs, $\theta = 360°$, at 10 μs it becomes 3600°, or in other words, a linear accumulation in phase difference. The output of the phase detector is a ramp going from zero volts towards infinity, as shown in **fig. 2A**. The VCO can be considered a perfect integrator.

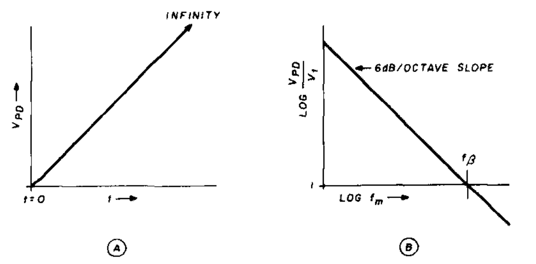

fig. 2. Step and swept frequency response of VCO-phase detector combination.

Since θ, and therefore V_{PD}, are a function of time and V_t, they must also be a function of signal frequency and amplitude at V_t.

Fig. 1B, is similar to **fig. 1A** except that in **fig. 1B**, the tune line is connected to a swept frequency sinewave source with fixed amplitude (in this case 1 volt P-P). Plotting the ratio V_t/V_{PD} versus f_M (**fig. 2B**)

By Craig Corsetto, WA6OAA, 4312 Marlowe Drive, San Jose, California 95124

shows a 6 dB/octave slope and crosses unity ($V_t = V_{PD}$) at a frequency:

$$f_B = \frac{K_{VCO} K_\phi}{2\pi}$$

$$f_B = \frac{(1\ MHz/volt) \cdot (5.73v/RAD)}{2\pi}$$

$$= 912\ kHz$$

Besides having a frequency response similar to a perfect integrator, it also has a constant 90 degree phase lag when comparing V_{PD} to V_t . This can be seen by using fig. 1B and fig. 3. At time A the voltage on V_t is zero volts (f_{VCO} in phase with f_{REF}).

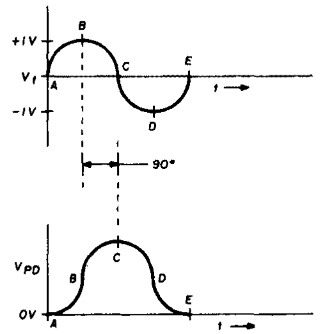

fig. 3. Plot of V_t and V_{PD} showing 90 degree phase lag due to integration of phase difference between f_{VCO} and f_{REF}.

Therefore, $V_{PD} = 0$ *volts*. As voltage V_t approaches +1 volt at time B, f_{VCO} becomes greater than f_{REF} so the phase difference at the phase detector accumulates causing V_{PD} to rise. Between time B and C the voltage on the VCO tune line decreases, but because f_{VCO} is still greater than f_{REF} the accumulated phase difference continues to increase, only at a decreasing rate. A plot of both V_{PD} and V_t over 360 degrees shows that V_{PD} lags V_t by 90 degrees.

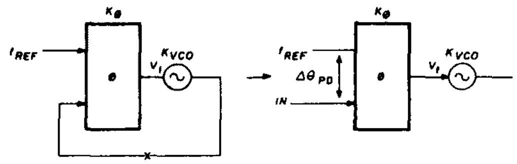

fig. 4. Breaking a simple loop to examine open loop properties.

open loop properties

Gain, frequency, and phase are open loop characteristics as shown in fig. 4. The total gain, K_t, is

$$K_t = K_\phi K_{VCO}$$

This means that with a phase difference at the input of the phase detector ($\Delta\theta_{PD}$) there is a frequency change of the VCO frequency (Δf_{VCO}), or:

$$K_t = K_\phi K_{VCO} = \left(\frac{\Delta V_t}{\Delta\theta_{PD}}\right) \cdot \left(\frac{\Delta f_{VCO}}{\Delta V_t}\right)$$

$$= \frac{\Delta f_{VCO}}{\Delta\theta_{PD}}$$

frequency divider

Adding a frequency divider (fig. 5) with gain term, K_N, changes the open loop gain equation to:

$$K_t = K_\phi K_{VCO} K_N$$

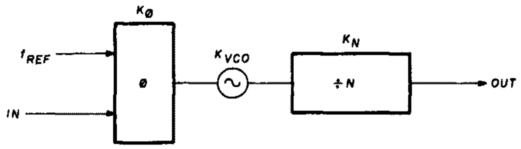

fig. 5. Open loop with divide by N.

Plotting K_t as a function of f_M gives a plot similar to that of fig. 2B except unity open loop gain occurs when:

$$f_M = \frac{K_\phi K_{VCO} K_N}{2\pi}$$

loop filter

Because of the more complex gain and phase versus frequency characteristics of the loop filter, K_F, it is examined separately, and then added to the loop.

Use of a simple loop filter (fig. 6) provides the gain and phase plots (fig. 7). Gain equals unity when:

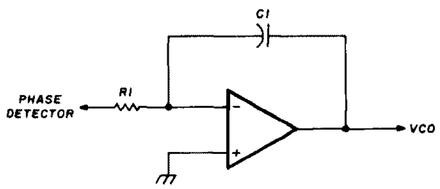

fig. 6. Simple op-amp integrator as a loop filter.

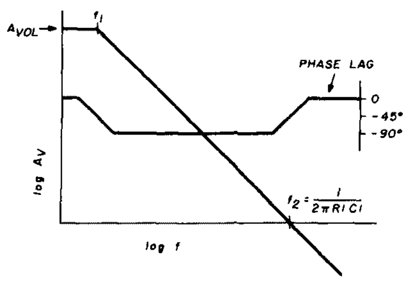

fig. 7. Amplitude and phase plot of simple loop filter in fig. 6.

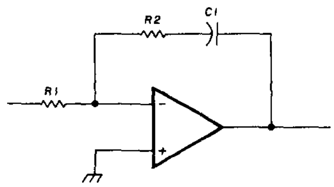

fig. 8. Adding resistor R_2 to integrator for better control of gain and phase.

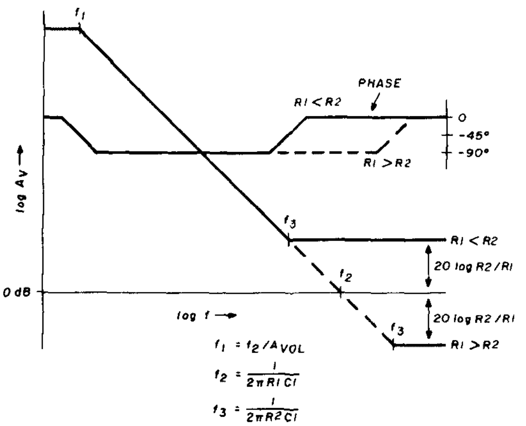

fig. 9. Amplitude and phase plot of loop filter in fig. 8.

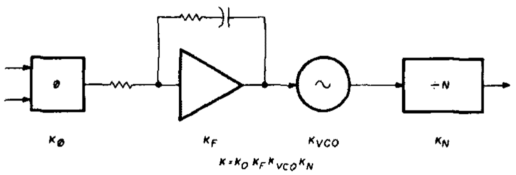

fig. 10. Open loop with loop filter.

$$f = \frac{1}{2\pi R1C1}$$

The phase will lag by 90 degrees from approximately f_1 to f_2.

To have more control over gain and phase, another resistor, R2, is added to the loop filter (**fig. 8**). The effects of R2 are seen in **fig. 9**.

open loop phase

By adding K_F to K_ϕ K_{VCO} K_N, (**fig. 10**) the final open loop characteristics are determined. This is accomplished by plotting K_F and K_ϕ K_{VCO} K_N separately on log-log paper and *then adding the gains* in dB together (**fig. 11**).

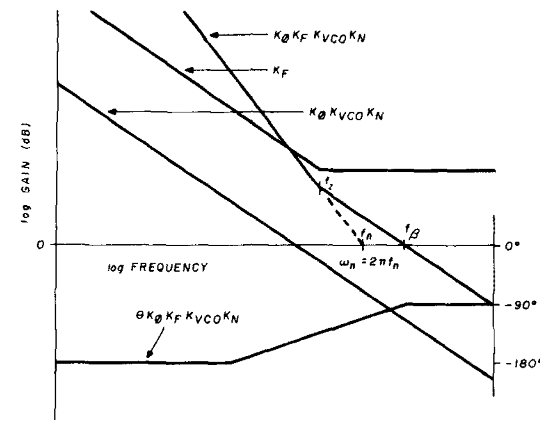

fig. 11. Plotting K_F separately from K_ϕ K_{VCO} K_N and then adding them to show combined frequency and phase characteristics.

The plot of phase shows the additive phase lags of both the loop filter and the VCO. Though the phase lag from the VCO is constant, the loop filter will start to reduce its own phase lag at:

$$f_Z = \frac{1}{2\pi R2C1}$$

About one decade of frequency past f_Z the loop filter phase lag will be close to zero.

closing the loop

Finally, closing the loop gives an essentially flat response from dc to frequency f_β, which is the frequency where the open loop gain equals unity (**fig. 12**). From frequency f_β and higher the closed loop gain follows the open loop gain. f_β is defined as the closed loop 3 dB bandwidth.

Since the PLL uses *negative* feedback, 180 de-

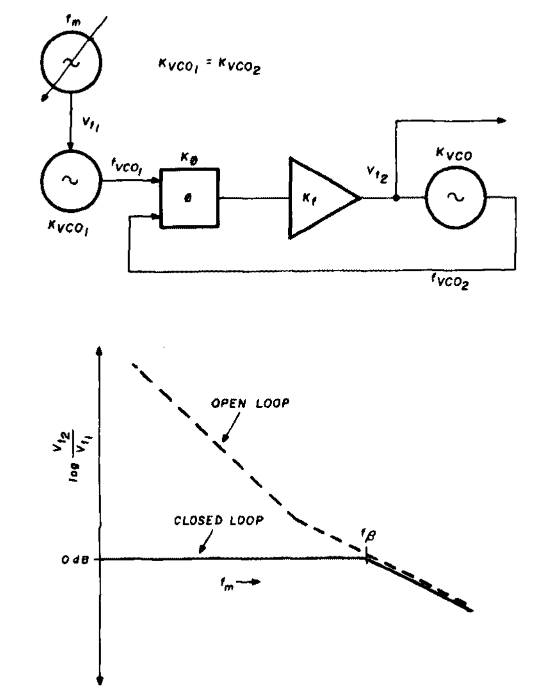

fig. 12. Open and closed loop responses of a simple loop. (Oscillator f_M is the modulating f_{VCO_1}. Both K_{VCO_1} and K_{VCO_2} are equal.)

grees of constant phase shift is required in the loop. This is usually obtained through proper selection of phase detector inputs. However, since both the loop filter and the VCO introduce phase lag, and if the accumulated lag becomes 360 degrees before the frequency where unity (open loop) gain occurs, f_β, the loop will oscillate. Since the phase detector introduces a constant 180 degrees, and the VCO a constant 90 degrees, the loop filter must introduce less than 90 degrees phase shift at f_β. The closer the loop filter comes to 90 degrees the more the loop will be inclined to ring to the point where it finally oscillates.

The distance the total loop phase lag is from 360 degrees at frequency f_β is called *phase margin* with zero being the smallest distance and 90 degrees the maximum. The damping factor, ξ, is related to phase margin by:

$$\text{phase margin} = 90^\circ - \tan^{-1}\sqrt{\frac{\sqrt{1 + \frac{1}{4\xi^4}} - 1}{2}}\ ^{*}$$

The loop natural frequency, ω_n $(= 2\pi f_n)$, sometimes called the loop resonant frequency, is the frequency the loop would oscillate at if $\xi = 0$ (no damping). While ω_n has limited importance in basic PLL synthesizer design, it is very important in other PLL uses such as PLL demodulators. In this article, however, ω_n is used only as an intermediate answer in calculations.

phase noise

Phase noise is random phase fluctuations of a signal. Typically hard to see on an oscilloscope except in severe cases, phase noise is more often examined on a high-resolution spectrum analyzer as shown in **fig. 13**. Since an identical noise component exists on either side of the carrier, usually only a one-sided spectrum, SSB, is shown on phase noise plots (**fig. 14**). Noise power is usually normalized to a 1 Hz bandwidth.

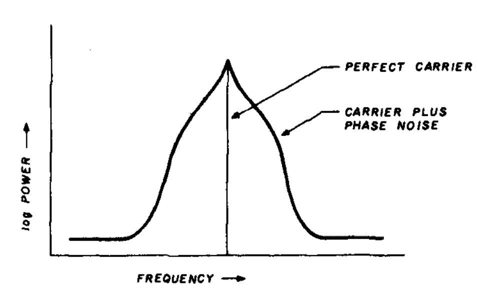

fig. 13. Oscillator phase noise as seen on a spectrum analyzer.

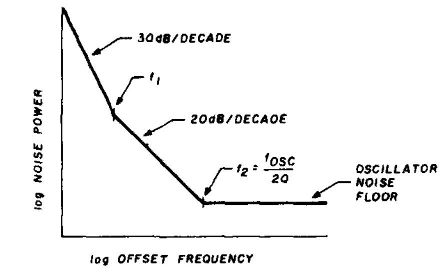

fig. 14. SSB plot of phase noise on a carrier using log-log scales showing effects of resonator Q and l/f noise.

Fig. 14 shows a typical noise spectrum of an oscillator. The phase noise will start to rise from the oscillator noise floor at a 20 dB/decade slope at:

$$f_2 = \frac{f_{OSC}}{2Q}$$

where Q is the loaded resonator Q of the oscillator.

*William F. Egan, *Frequency Synthesis by Phase Lock*, John Wiley & Sons, 1980, page 46.

At small frequency offsets from the carrier the noise rises with a 30 dB/decade slope due to additive 1/f noise generated by the oscillator active device.

In high performance receivers not only are synthesizers required to assume crystal-like frequency stability but crystal-like phase noise performances as well. PLL frequency synthesizers have the unique ability to improve oscillator (LC, RC, etc.) phase noise performance approaching that of the reference oscillator (typically crystal).

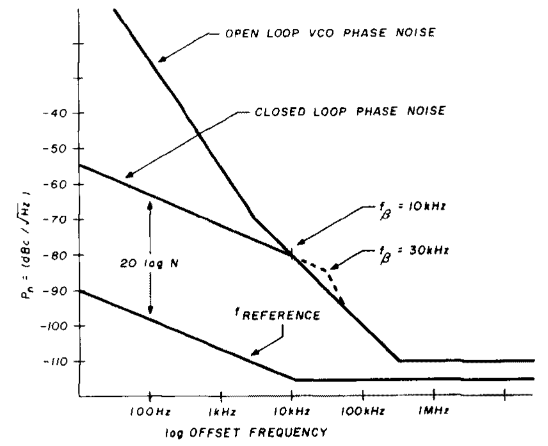

fig. 15. SSB phase noise plot comparing closed and open loop VCO phase noise and the effect of N.

Fig. 15 shows an open loop phase noise plot of a VCO. Also shown is what might be the phase noise of a crystal controlled reference oscillator. When the VCO is incorporated in a PLL and the loop is closed the VCO achieves improved phase noise performance within the bandwidth of the loop, and assumes its normal open loop phase noise response outside the loop bandwidth. However, the phase noise within the loop bandwidth is largely determined by the value of N and the phase noise of f_{REF}. This can be illustrated by a simple loop where the divider equals one, and the reference oscillator suddenly incurs a short term shift in frequency (phase noise). The phase detector senses this and generates a voltage to adjust the VCO frequency to the new reference oscillator frequency.

When the programable divider value, N, is greater than one, then in order for the phase detector to see zero phase difference at its inputs, *after* the same short term shift in f_{REF}, the VCO must change to a frequency N times the change in f_{REF}. Therefore the effect of N on VCO phase noise, within the loop bandwidth, is to multiply it by N, showing up on phase noise plot as a 20 log N more phase noise than the reference oscillator. This effect of N on phase noise can be a determining factor when selecting the optimum loop bandwidth (see **fig. 15**).

Often in medium performance receivers it is acceptable just to have the local oscillators meet the phase noise performance of the VCO. It happens that this can simplify the synthesizer design, allowing N to be large and f_β small for optimum phase noise performance, instead of using a more complex design in order to keep N small and loop bandwidth wide for improved phase noise performance. Problems encountered in using the simpler loop is an increase in lock time (due to the narrower loop bandwidth) and increasing difficulty in reducing reference feedthrough due to f_β approaching f_{REF}.

For increased synthesizer phase noise performance without too high a lock time and reference feedthrough, a more complex loop structure is required. **Figs. 17-19** show more complex loops, with the simple loop in **fig. 16** as a reference, while **fig. 20** plots their approximate relative phase noise performance.

The loop in **fig. 17** requires closer attention. This loop used in-loop mixing to reduce the required size of N. With N reduced the phase noise within the loop bandwidth decreases and the optimum value of loop bandwidth increases, decreasing lock time. More importantly, the local oscillator used in the in-loop mix-

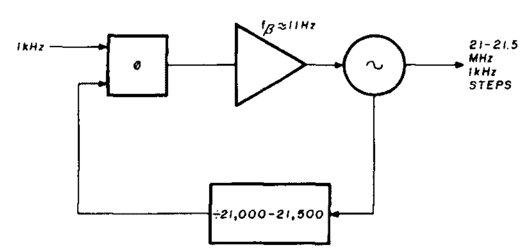

fig. 16. Simple frequency synthesizer to be compared against those in fig. 17-19.

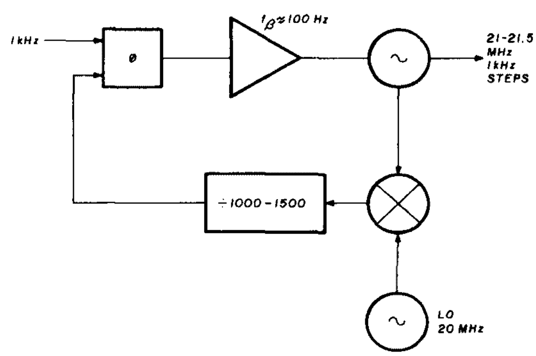

fig. 17. More complex synthesizer to reduce value of N and therefore phase noise within f_β.

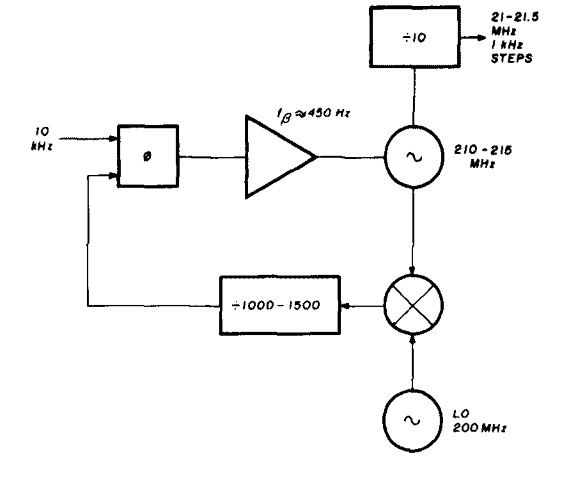

fig. 18. A synthesizer similar to than in fig. 17 only using a ÷ 10 on the VCO output to reduce frequency by 10 and phase noise by 20 log 10.

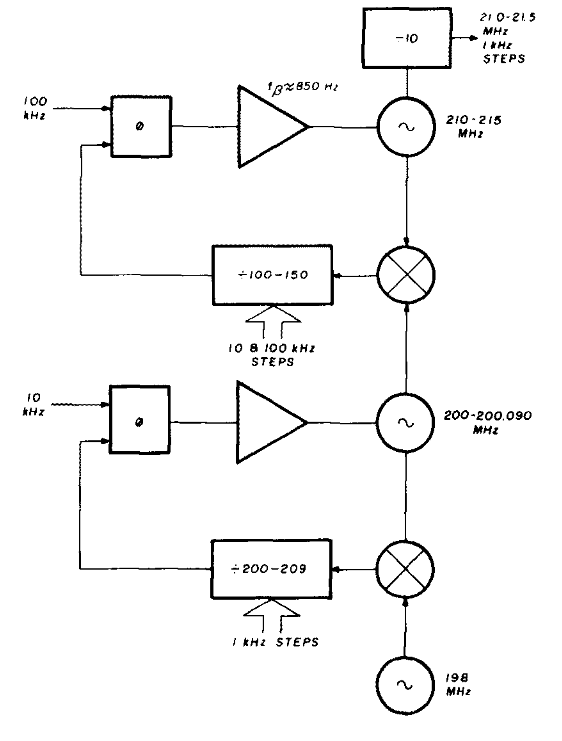

fig. 19. Using another synthesizer for in-loop mixing to help improve phase noise performance. Using various adaptations of this can help keep lock time short and phase noise low.

ing can be a VFO, VCO, VCXO, or even another synthesizer (**fig. 19**) for finer frequency steps. Some Amateur radio transceivers use this technique by letting the synthesizer step in 1 kHz increments by changing N and then step in 100 Hz increments by tuning the VCXO local oscillator using a DAC (Digital to Analog Converter).

VCO phase noise and K_{VCO}

In real synthesizers the phase noise is more than expected due to additive noise generated by both active and passive devices modulating the VCO. Therefore, for best performance, choose the components used in the synthesizer wisely, especially those in and around the VCO and loop filter.

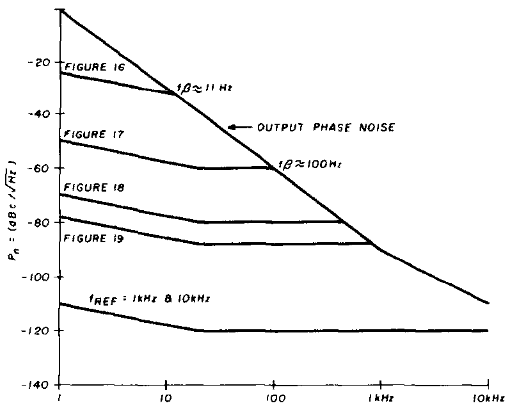

fig. 20. Relative phase noise plot of synthesizer in figs. 16-19.

The effects of noise and spurious signals on the VCO tune line can be approximated if the modulation index caused by these is known to be small.

$$P_n \approx 20 \log \frac{2\sqrt{2}\, V_t K_{VCO}}{8\pi f_M}$$

where P_n = SSB noise power in a 1 Hz bandwidth, or spur power, in dBc

V_t = RMS noise, or signal voltage on VCO tune line

f_M = frequency of noise or signal on VCO tune line

For example a VCO with a $K_{VCO} = 628.3\ (10^3)$ has 100 μv (RMS) of spurious signal at 1 kHz on its VCO tune line. This spurious signal appears as 1 kHz sidebands with a sideband spur power of:

$$P_n = 20 \log \frac{2\sqrt{2}\ (100\ \mu v)\ 628\ (10^3)}{8\pi\ (1\ kHz)}$$

$$or\ -43\ dBc.$$

frequency division and phase noise

External to the phase locked loop, synthesizer phase noise can be reduced through the use of frequency division by 20 log N if a reduction of frequency by N is acceptable. This is quite common in commercial synthesizers and is shown in **figs. 18** and **19**.

As is always the case, using this technique to reduce phase noise has its limits, in this case the noise floor of the digital divider.

For MOS, TTL, and ECL devices, the noise floors are typically -120 to -140 dBc $\sqrt{Hz}$, MOS devices having poorer noise performance than TTL and ECL.

In the third and final part of this series, two techniques for designing a 5.000-5.500 MHz synthesizer — as well as trade-offs common in PLL design circuits — will be discussed.

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ham radio

10-GHz weather radar

A unique application for a Gunnplexer transceiver

Commercial weather radar systems have become quite popular in recent years and have proven their true life-saving ability. And over and above their use during dangerous storms, they have now become almost a part of our daily lives. The only drawbacks to these commercial radar systems are their inability to scan a small area (such as an individual's neighborhood), and the fact that commercial radar scans are generally made available only at fixed times.

To overcome these limitations, I have designed a mini weather radar system which is both inexpensive and easy to build. A short scanning range was my only criterion. Guidelines are provided for those interested in pursuing and refining this unique concept.

This mini weather radar is based on two simple facts: a radio signal travels one mile in 5.4 microseconds; and an electron beam can be timed to travel across a CRT's screen at a predetermined rate. If a radio wave leaves a transmitting antenna at the same time that a horizontal line begins on the left side of a CRT display, that radio wave will travel approximately 12.1 miles in the 63 microseconds it typically takes for one scan of a horizontal line in conventional TV.

operation

The first version of this unit is shown in **fig. 1**. An oscilloscope with horizontal trigger output is used with a 10-GHz Gunnplexer, an NE555 pulse modulating circuit, and a high gain amplifier to produce a distance-and-azimuth Amateur weather radar. The Gunnplexer's mixer uses the unit's constantly transmitted signal for a local oscillator, eliminating the need for a T/R device. A horizontal scan-initiating sync pulse from the oscilloscope's horizontal trigger output is used to key the NE555 oscillator, which becomes the modulating signal for the Gunnplexer's varactor input.

The Gunnplexer transmits its signal toward a distant object, such as heavy storm clouds, and its reflection is then received. The return signal heterodynes in the Gunnplexer's mixer, the output of which drives a high-gain amplifier. The amplified echoes are then fed to the oscilloscope's vertical input for display. Since the scan started during the pulse-transmission time, its round-trip delay time is indicated in **fig. 2** as displacement on the screen.

The Gunnplexer's modulating frequency should be between 200 kHz and 1 MHz. This frequency deter-

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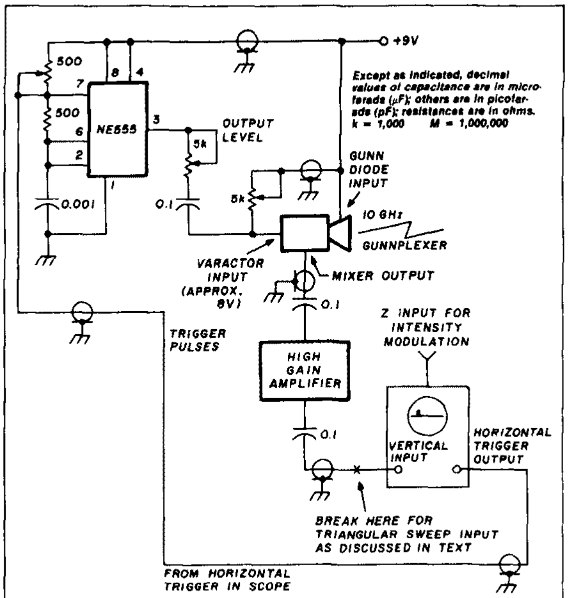

fig. 1. Arrangement of the first-generation Amateur weather radar, using an oscilloscope for generating timing pulses and displaying returned echoes.

mines the exact pulse width. Using the formula $F = 1/T$, a 200-kHz modulating frequency translates into approximately 5 microseconds: approximately one mile on the screen's display. A 1-MHz modulating signal encompasses approximately 1 microsecond (1/5 mile) on the screen. Although the Gunnplexer transmits a carrier continuously, only the brief modulating pulse causes indications on the CRT display. This unit will provide distance indications and information on the density of rain clouds, but it does not provide information on the direction of storm clouds. This shortcoming does not usually pose a problem.

refinements

The second version of this weather radar uses an NE555 triangular wave generator and adds a modified oscillating house fan motor to make possible directional (azimuthal) readings. An outline of this arrangement is shown in **fig. 3**. The fan motor causes side-to-side movement of the remote Gunnplexer, while a microswitch at one end of the sweep provides synchronizing pulses for the NE555. The output from the triangular wave circuit is applied to the oscilloscope's vertical input; the high-gain amplifier's output is applied to the intensity (Z) input of the oscilloscope. The oscilloscope is adjusted for a dark screen until an intensity-modulating echo is received. Range markers can be added to this system with another NE555 circuit, or can simply be drawn on the screen with a washable-ink felt pen. I also suggest that you replace the oscilloscope's P-1 phosphor cathode ray tube with a long-persistence-display P-7 equivalent tube.

During operation, horizontal sweep pulses from the oscilloscope trigger the NE555 modulation circuit, while vertical sweep is provided by the fan-motor-synchronized triangular wave generator. The Gunnplexer transmits a 200-kHz-modulated 10-GHz signal which is reflected according to cloud density and displayed on the screen. Horizontal displacement on the CRT indicates distance, and vertical height (with respect to the base line) indicates direction.

The received echoes are heterodyned with the Gunnplexer's carrier, amplified, and applied to either the oscilloscope's Z modulation port or vertical input (depending on the particular version of weather radar you build). Although a low-power 10-GHz Gunnplexer can transmit and receive over line-of-sight distances

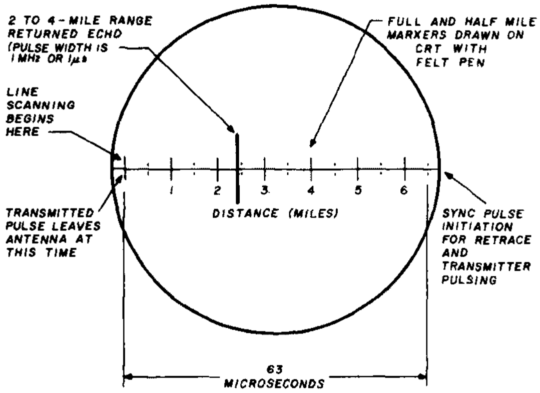

fig. 2. Visual display obtained with first-generation Amateur weather radar. The 10-GHz Gunnplexer transmits a modulation pulse at the same time a scan line begins on screen at left. The time periods shown provide a range of approximately 6 miles.

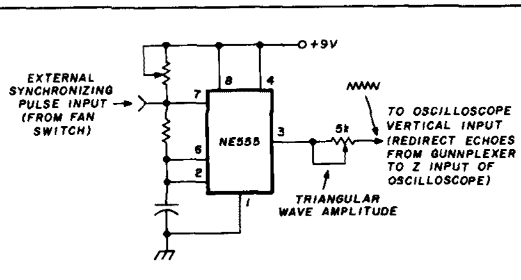

fig. 3. Triangular wave generator for vertical deflection in synchronization with Gunnplexer movement, as described in text. Circuit insertion point is shown in fig. 1.

in excess of 20 miles, I suggest that you limit its range in this application to less than 12 miles (a number chosen because of the oscilloscope's horizontal scan rate and the frequency of the Gunnplexer's modulator). This safety margin ensures reliable operation and permits accurate interpretation of the displayed information.

a third mini radar

A third version of the mini weather radar is being developed now, and its capabilities look promising. This system employs a used black-and-white TV set rather than an oscilloscope, using the TV circuits to generate precisely timed transmitter pulses and to display the reflected signals. Since the set's tuner and i-f stages are bypassed, with the radar information applied directly to the video amplifier section, usable TVs could be picked up inexpensively at TV repair shops.

An outline of the system is illustrated in **fig. 4**. Narrow-width (and highly accurate) horizontal AFC pulses obtained from one of the low-voltage windings on the TV's horizontal output transformer trigger the Gunnplexer's NE555 modulator. This arrangement establishes radar timing while maintaining horizontal line sync in the television. A single-shot multivibrator or Schmitt trigger proves useful for buffering flyback pulses. Alternatively, a dropping resistor may be placed in the keyed AGC line for lowering that voltage to approximately 5 volts. A schematic diagram of the TV is needed for locating the keyed AGC line and determining its voltage level (don't "hunt around" in the horizontal output section: HIGH VOLTAGE!). The desired modulator-keying line is usually found connected to the AFC's discriminator diodes. You will also see a "second line" carrying sync pulses from the sync separator stage in the AFC's discriminator. That guidepost will help you locate the desired AGC keying takeoff point.

Output from the Gunnplexer's modulator is applied to that unit's varactor. The returned echoes are amplified to approximately **4** volts by a wideband amplifier similar to a single video stage in a television set, and applied to the TV's video amplifier section (a suitable injection point for this signal is quite often between the contrast control wiper and ground. If an ac/dc TV is used, be sure this point is not hot). The resultant echoes are displayed as intensity variations on the screen, with distance and timing calibrations provided by techniques similar to those in the second unit, described earlier. The next step in this system's development will include electronically sweeping the 10-GHz radar signal in sync with the TV's 60-Hz vertical scan rate. An illustration of the resultant display is seen in **fig. 5**.

A fourth generation Amateur weather radar is presently on the drawing board, and this version holds some truly exciting possibilities. Basically, this system employs an inexpensive home computer, such as the Apple II or TRS 80C, for generating all the required timing pulses and for creating a color-level display similar to that of commercial units. Since these microcomputers are easily programmed for various timing ranges and color presentations, they are quite useful for Amateur radar work. The integration of A-to-D and D-to-A converters in these units makes computer interfacing relatively simple.

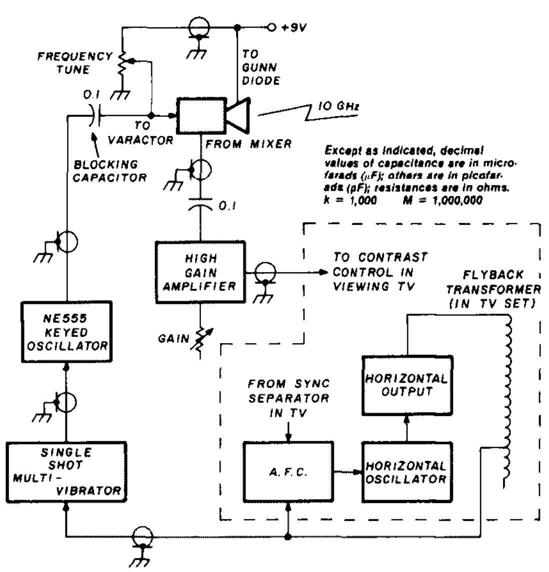

fig. 4. The third-version Amateur weather radar uses a modified television set for timing and display of target information.

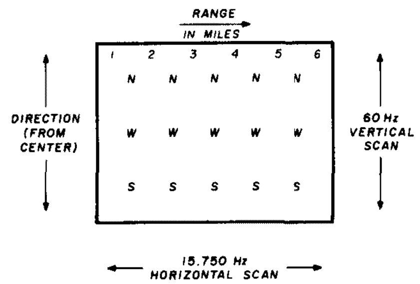

fig. 5. Display of sector scan as accomplished by version-three Amateur weather radar. A modified TV set is employed.

Alden Electronics (Washington Street, Westborough, Massachusetts 01581) manufactures a weather chart recorder (Model 9321), available in assembled or kit form, that displays data generated by NOAA and other sources. — **Editor**

ham radio

installing effective ground systems

A few hours' work improves signal, reduces RFI

Many Amateurs strive for maximum efficiency from their antennas yet fail to give the same consideration to their ground systems. All too often their ground consists of a metal rod or pipe driven into the soil at some convenient point. Any handy piece of wire is run between it and the rig. Many times, the transceiver or transmitter is the only piece of equipment connected to the ground. While this type of wiring practice may be satisfactory for ac or dc, it is totally ineffective at radio frequencies.

Poor grounding can worsen TVI and RFI, reduce the effectiveness of the shielding in your rig and allow rf feedback into various pieces of equipment. Good grounding can noticeably enhance the performance of antennas on 160, 80; and 40 meters.

There are two phases in the establishment of an effective ground system. The first involves properly bonding together all equipment in the shack. The second entails the construction of a low resistance earth ground.

what should be bonded?

All pieces of equipment in the shack should be correctly bonded together. This includes transceivers, transmitters, receivers, power supplies, keyers, antenna tuners, etc. Unlike dc or ac, all current flow is at or near the surface of a conductor at radio frequencies. Thus, the ideal bonding material should have a large surface area to present the least impedance to the flow of rf currents. The best and most expensive conductors are braided copper strap and flashing copper, which both possess large surface areas. No. 6 gauge copper wire, normally used for grounding, has *less* surface area than the shield of RG-58 coaxial cable. In addition to being expensive, large gauge solid copper wire is very stiff, making it difficult to use.

A less expensive substitute for copper flashing or braided copper strap is the shield of RG-8 coaxial cable. It isn't necessary to use new cable. If you replace your coaxial cable every three to five years, as do most hams, the cable you replace will probably be adequate for grounding purposes.

A ground bus using the entire cable is shown in **fig. 1**. It is not necessary to pull the shield off the coaxial (a nearly impossible task). The center conductor is not used in order to avoid making the cable self-resonant. In the installation shown in **fig. 1**, short lengths of RG-58 cable were used to make connections to individual pieces of equipment. Again, only the shield was used. To facilitate connection, the center conductor and insulation were cut to allow 2 inches of free shield at each end. The RG-8 bus was stapled across the back of the desk to place it as close as possible to the equipment.

cold water pipe

Amateurs are often advised to ground their equipment to the nearest cold water pipe. But this is impossible if the pipe is plastic or corroded metal. Even with copper pipe, these makeshift grounds may be difficult or impossible to reach — unless you operate from the kitchen, bathroom, or laundry room. The use of water pipe instead of a good earth ground is justified only if you live above the first floor of an apartment building.

achieving good earth grounds

The construction of a good earth ground, while not difficult, involves more than driving a metal rod into the lawn. The soil into which the ground rod is driven should present minimum resistance to the flow of electric currents. When completely dry, most soils are non-conductors. Pure water is a very poor conductor. The basic electrical conductivity of a soil is the result of electron transfer through electrolytes dissolved in the water present in that soil. However, there are many factors that can affect this basic conductivity.

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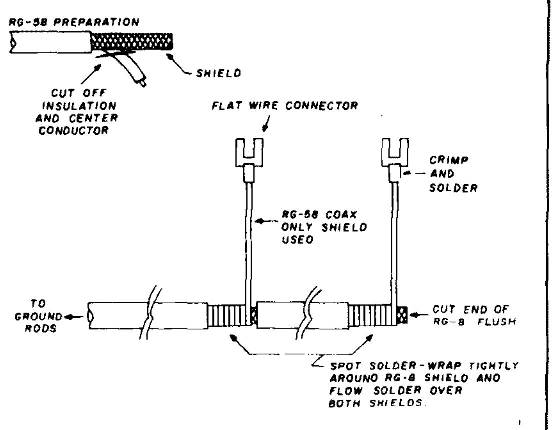

fig. 1. Using RG-8 coaxial cable as a ground bus.

material	conductivity (millimhos/meter)
poor soil	1-5
average soil	10-15
very good soil	100
water	10-15
salt water	5000

fig. 2. Soil conductivity chart.

Soil type has an important effect on local ground resistance as shown in **fig. 2** and **3**. Commercial broadcast stations require excellent ground systems for their vertical antennas. This is why they show a marked preference for low swamps over rocky mountain tops. Soil particle size and density influence conductivity, which explains the differing resistances between clay and sandy gravel. A fact to consider in locating your own earth ground is that soil type can change over very short distances, both vertically and horizontally. The flowerbed may look like a good place for a ground rod, but it won't be if the topsoil is resting on gravel or rock.

The moisture content and concentration of dissolved salts markedly affect soil resistance (see **figs. 4** and **5**). There is a direct correlation between increased *soil* moisture and *lowered* resistance. The same effect is noted with respect to dissolved salts in the soil. Little information is available on the changes of soil conductivity with respect to different types of salts. The importance of soil moisture and salt concentration cannot be overemphasized. Salts disassociate into the ions involved in electron current flow; water provides the medium to facilitate easy flow of these ions through the soil.

Two other interrelated factors affect soil conductivity. As shown in **figs. 6** and **7**, soil resistance rises quickly as the temperature drops. Soil conductivity is lowest during the winter months and highest in summer. However, these temperature and climatic effects become less noticeable with increasing depth in the soil. Those of you living in cold climates will notice the large jump in soil resistance as the soil freezes. This is due to ions being trapped in the crystalline structure of the ice. Maintaining a low ground resistance during winter months means driving your ground rods well below the frost line.

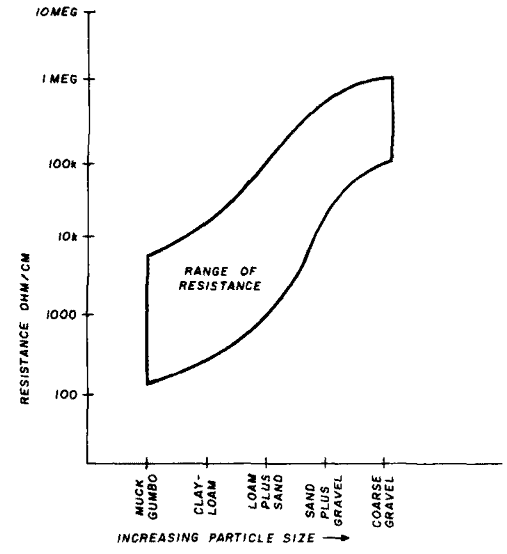

fig. 3. Resistance values of various soils.

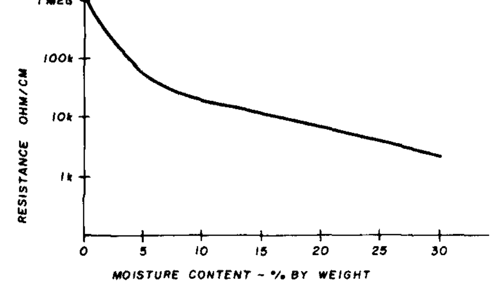

fig. 4. Effects of moisture in sandy loam soil.

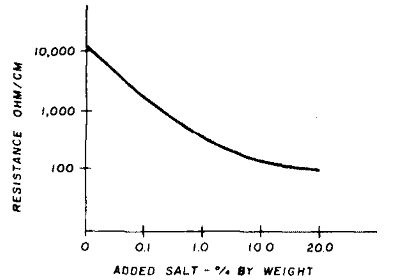

fig. 5. Effects of dissolved salts in sandy loam soil.

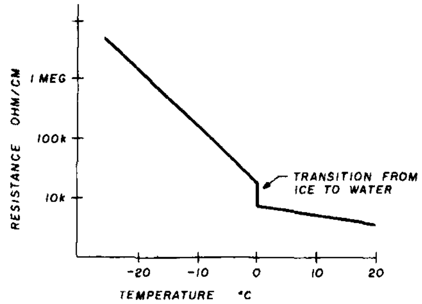

fig. 6. Temperature affects on sandy loam.

chemical treatment

Chemical treatment of the soil around a ground rod will increase the effectiveness of your ground system.* Rock salt, copper sulphate or magnesium sulphate (commonly known as Epsom salt) will inject large quantities of ions into the soil increasing its conductivity while reducing seasonal variation. However, these salts will gradually be washed away by rain and groundwater. They should be replaced every one to three years depending on climate and soil type. Rock salt is the most readily available and least expensive. Magnesium sulphate (Epsom salt) is available at any drugstore. Copper sulphate is the most effective but the most expensive.

ground rods

Copper pipe makes the best ground rod, but unfortunately, it's expensive and too soft to be driven into the ground. The best practical rods are made of steel plated with copper to decrease resistance. These are available through most Amateur Radio outlets. Remember that the length of a ground rod is more important than its diameter. Doubling its length will cut resistance by 40 percent, while doubling the diameter will result in only a 10 percent reduction.

Spaced rods provide large reductions in the resistance of your earth ground system as shown in **fig. 8.** Ground rods should be separated from each other by a distance equal to their length. This reduction of resistance is not proportional to the number of rods in the system. Three rods spaced 15 to 25 feet apart will provide an optimal ground for all Amateur stations.

A typical ground rod set-up is shown in **fig. 9.** This is most effective since the chemical salts are distributed through a large volume of soil. Unfortunately, these salts are toxic to vegetation and can leave a ring of bare soil around each rod.

An alternate installation method for ground rods is shown in **fig. 10.** This was used for my station since I have a small city lot and all the good locations for driving ground rods seemed to be covered with petunias. All excavation was done with a post-hole digger to minimize damage to the flower beds. While not quite as effective as the method in **fig. 9**, it allows much greater latitude in the placement of your ground rods.

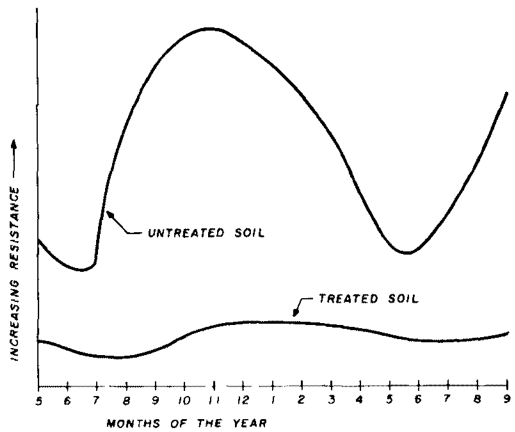

fig. 7. Seasonal variation of ground resistance.

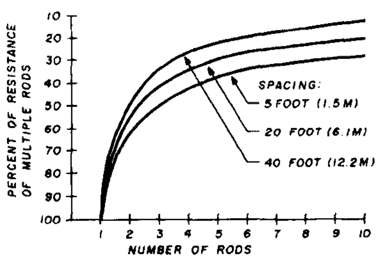

fig. 8. Resistance reduction of multiple rod ground over single rod ground.

*Use extreme caution in applying chemicals that may pollute ground water, wells, or soil. **Editor.**

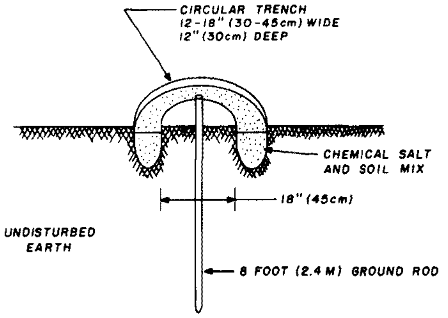

fig. 9. Typical ground rod set-up.

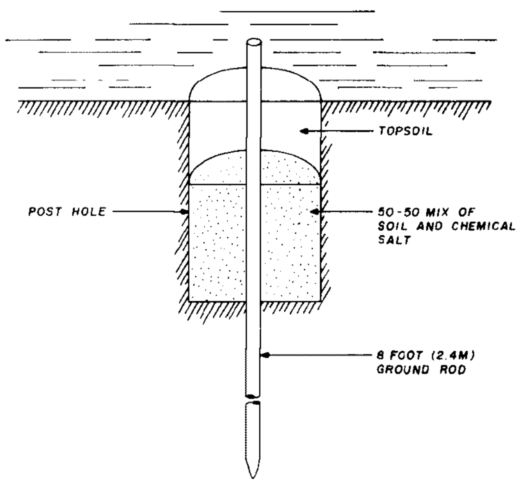

fig. 10. Ground rod with treated soil for confined areas.

The proper construction of a low resistance earth ground system is a relatively simple task. It can eliminate the problem of "rf in the shack," reduce your TVI potential, and improve the performance of low band antennas — not a bad trade-off for several hours of work.

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ham radio

locate orbiting satellites

The motion of satellites around the earth is explained to help you find and use Amateur satellites

For the beginning satellite chaser, probably the most perplexing question is, how do I find out when I can hear it? Although there are now numerous ways of finding the answer to that question, the newcomer — and possibly some old-timers — will find it very helpful to understand the spatial relationships that exist between satellites and the earth. With the aid of a simple globe of the earth and some basic information on satellite behavior, a clearer picture of how to find satellites emerges.

Look at a globe of the earth and visualize it divided exactly in half by a flat planar surface. Imagine this plane at a slight angle to the polar axis, so that it passes to one side of the north pole and to the other side of the south pole. Now consider this plane to be stationary, with the earth rotating freely through it. Note that twice during each rotation each point on the earth's surface passes through this plane — except for small regions near each pole. At each passage the plane will appear to approach the earthbound observer from the east and disappear to the west, just like the sun and moon.

satellite paths

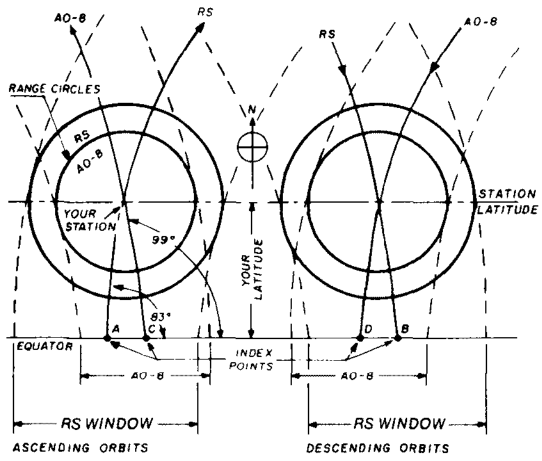
What does this have to do with satellites? Satellites travel around the earth in a path that stays on a flat plane of this kind, and the planes of the various satellites are completely independent of one another. Much of the apparent movement of the satellites is merely the movement of the earthbound observer being carried toward and away from the orbital plane.

The problem of discovering when a given satellite comes within useful distance is then really the problem of finding out where the orbital plane is when the satellite is near your latitude. Since this information is different for every station, a convention is used that identifies the time and location of the satellite's equatorial crossing. From that information you can derive the specific data you need for your own location.

A satellite circling the earth in the orbit previously described crosses the equator twice during each revolution around the earth. The convention for reference data identifies only the northbound equator crossing. Information on the longitude and time of the first northbound equatorial crossing for the UCT (Greenwich) day is published and broadcast by ARRL, AMSAT, and others. In addition to this information, we must also know certain other facts about the satellite's orbit. The time it takes a satellite to revolve exactly once around the earth is called its *period*. The angular distance, measured at the equator, between successive northbound equator crossings is called the *increment*; measured in degrees of longitude, and mainly related to the earth's rotation. It is also important to know the *inclination angle*, the angle between the orbital plane and a plane through the earth's equator. All these orbital characteristics are unique to each satellite and are published in various Amateur magazines.

For the ham who really wants to get serious about tracking satellites, there is another item of information I can recommend. That is something that's not published but must be calculated by the individual for his own location. I have called it the *index point*, and I define it as the longitude of the intersection of the orbital plane with the equator that causes the satellite to pass directly overhead. These index points are different for each satellite, and, in fact, there will be two: one for northbound satellite movements (ascending orbits) and one for southbound move-

By John L. Hill, WØZWW, 2838 Lake Boulevard, North St. Paul, Minnesota 55109



The "windows": Satellites will pass within range of your station if they cross the equator inside your window. Points A and B mark the center of the windows for RS birds, points C and D for OSCAR 8. Northbound equator crossings will be farther west from points D and B, which mark the window centers for southbound crossings by 180 degrees minus ½ increment. At stations located on the equator, the window widths are 75 degrees for RS birds and 58 degrees for OSCAR 8. For station locations at other latitudes, the window will widen by the reciprocal of the cosine of the station latitude (up to about 50 degrees either north or south). NOTE: Stations in the southern hemisphere should invert this figure and interchange the terms "ascending" and "descending."

fig. 1. "Window" determination for OSCAR 8 and RS satellites.

ments (descending orbits). Let's take a closer look at these index points.

If the orbital plane had an inclination angle of 90 degrees (perpendicular to the equator), one might predict that the index point would have the same longitude as the ground station. That is not quite true, however, because it takes time for the satellite to move from the equator to the station and the rotation of the earth during this interval has moved the station eastward by one degree longitude for each four minutes. Thus the orbital plane must be positioned east of the station longitude to accommodate this fact.

Since the inclination angle of the orbital plane is not 90 degrees for any of the satellites now in orbit, there will be a difference between the longitude of the equator crossing point and the ground station because of this "leaning" of the plane. An effective way to visualize this is to stretch a rubber band *around* the globe, making a great circle. Position this great circle so that it passes through the ground station location, and make the northernmost and southernmost points tangent to latitude parallels equal to the inclination angle. For OSCAR 8 this angle is 99 degrees, so the great circle will come tangent to 81 degrees, passing to the left of the north pole (as you view your station's location) and to the right of the south pole. For the Russian satellites (RS-3 through RS-8) the inclination angle is 83 degrees, which makes the tangency points nearly the same, but on opposite sides of the earth's poles.

With the assistance of this great circle it's apparent that for inclination angles of less than 90 degrees, the plane crosses the equator slightly west of the ground station's longitude; for inclination angles greater than 90 degrees, it crosses slightly east. The contribution to the displacement due to the earth's movement is always toward the east; thus these two effects almost cancel one another for the RS birds. They add for OSCAR 8.

For ground stations with latitudes of not greater than about 60 percent of the inclined orbital plane's northernmost (or southernmost) latitude, it's possible to calculate the location of the index points without resorting to spherical geometry and still get accurate enough results for Amateur communications. For example, consider a ground station at 40 degrees north latitude and 80 degrees west longitude. And let us use the data for the RS-8 satellite, because it provides some easily understood calculations (see **fig. 1**). RS-8 has a period of just under two hours, an increment of almost exactly 30 degrees of longitude, and an inclination angle of 83 degrees. Its height above the earth's surface is 1680 kilometers, a value we will also need.

The first value we will need to derive is the displacement caused by the earth's rotation. Since the satellite travels at a velocity of 3 degrees per minute, it will take 13.3 minutes for it to travel north from the equator to the station's latitude. Meanwhile, the earth turned 3.33 degrees (13.3 divided by 4).

The *inclination factor* is the latitude multiplied by the cosine of the inclination angle. That is 4.9 degrees, but since it's west and the 3.33 degrees is east, they partially cancel, making the index point 81.47 degrees west longitude. The significance of this is that every orbit of RS-8 that ascends across the equator at 81.5 degrees will pass over the ground station 13.3 minutes later.

A similar index point can be found for satellite passes which ascend on the other side of the earth and descend over the station. The arithmetic for determining this index point has almost all been done, since the satellite now passes the station before it reaches the equator by 13.3 minutes. Both the previous displacements change direction, and the equator crossing on descent becomes 78.5 degrees west.

It's more useful to learn where the ascending equator crossing must be to produce the desired descending longitude. One might expect it to be exactly opposite (i.e. 180°), however, that is not correct since the earth has been turning during the intervening half-orbit. Therefore, the required longitude of ascension will be east of a point opposite by an amount proportional to the time required for this half-orbit transit. Since all satellite locations are expressed in longitude degrees west of zero, this point is most easily obtained by adding 180 degrees minus

one-half the increment. (Should the result be greater than 360 degrees, the correct value is obtained by subtracting 360.) For the example we've been using, this ascending crossing will be at 243.5 degrees west. It's useful to note that the 'bird' will arrive over the station 46.5 minutes after this crossing: one-half the period minus the time from the station to the equator.

So much for passes which are overhead. But what about the ones that come by without being that high in the sky? To find out when and where these occur, we must take up the question of range. When the satellite is just on our horizon (unless the horizon is obscured by a mountain or building) the ground station is at the 90-degree corner of a right triangle, the other two corners of which are the center of the earth and the satellite. The useful information to be drawn from this triangle is the angle between the lines joining the center of the earth with the other two corners. This angle defines the maximum range at which the satellite is visible on the horizon, and its value can be used to calculate the equator crossing longitudes of orbits that will carry the satellite inside a "range circle" that's centered on the ground station and has a radius of that dimension.

If the ground station were on the equator, we could merely add and subtract the range angle to the index points to derive the "window" through which every satellite would have to pass to come into ground station range. For ground stations off the equator — at latitudes greater than zero — the range window widens. Without getting into complicated spherical geometry, it's possible to estimate the width between eastern and western equator crossings at the range limits by dividing the range angle by the cosine of the station latitude, and adding and subtracting the resulting value to the previously calculated index points.

Using the RS-8 satellite and the ground station of the earlier example, the range angle is 37.7 degrees.* Dividing this by the cosine of 40 degrees widens this angle to 49.2 degrees. Adding and subtracting this gives limits of 32.3 degrees and 130.7 degrees, and provides a fair approximation of where ascending equator crossings will bring the satellite within range of this ground station. The same operation on the other index point gives a set of limits for descending orbits in terms of the longitude of ascension. For this station they are 194.3 degrees and 292.7 degrees west longitude.

The true boundaries of equator crossings are slightly beyond these calculated values, but the mathematics becomes too involved to make their determination worthwhile. Also, this simple scheme develops gross inaccuracies as the location of the ground station exceeds latitudes greater than about 60 percent of the maximum satellite latitude. Stations in higher latitudes need to use a different approach.

which satellites are in range?

Having established the edges of our own personal index windows, we now need a list of all the satellite equator crossings to see which of them will be in range. Lists showing all equator crossing longitudes and times for ascending orbits are available, but they are predictions, made months into the future, and tend to require corrections when the actual date arrives. Here is a simple technique, using an ordinary pocket calculator, for getting the same information from more correct reference orbit data broadcasts by W1AW and the AMSAT nets. If your calculator has a memory capable of storing an eight-digit constant, it can handle this procedure.

The idea is to combine the two values — time and longitude — into a single multiple digit value that can be repeatedly added to the initial reference data to yield successive orbital data. Again using data for RS-8, the satellite has a period of 119.7 minutes and an increment of 30.07 degrees. Combine these two values into a single expression and enter it into the calculator's memory. Make this value 199.70301 (why it's not 119.70301 will be explained later — see **calculator values**). Note that the increment has been rounded off to stay within eight digits.

Assuming that the reference data for RS-8 this day is 00:42:36 (hours:minutes:seconds) at longitude 260.4 degrees west, enter these two items as a single value (42.62604) and add the constant to get 242.32905. This can be mentally separated to be 2 hours, 42.3 minutes at longitude 290.5. Add the constant again and get 442.03206, 4 hours, 42.0 minutes at 320.6 degrees longitude. Add the constant once more and get 6 hours, 41.7 minutes at 350.7 degrees west longitude. The next addition comes to 841.43808, but since the longitude portion exceeds 360 degrees, merely subtract 0.036 and you find that this orbit ascends the equator at 8 hours, 41.4 minutes at 20.8 degrees west. You can continue to develop data for each successive orbit. Should the minutes value of the time come out between 60 and 99, merely subtract 40.0 and you have a correct value and can continue. In fact, you can even subtract 2400.0 when the time exceeds 2359 and continue on into the next day if you don't have new reference data for that day. If you go on too far, of course, the inaccuracies of rounding will build up.

*The angle is equal to arc cosine $[R/(R + H)]$, where $R = 6371$ km and $H = 1680$ km.

By comparing the longitudes of equator crossings with the limits you have computed for your two index windows, you can select those which will pass within the range of your station and establish the time that each will pass your latitude. The elapsed time from equator crossing to arrival at your latitude will be the same whether the equator crossing is such that the bird passes overhead or only out near the limits of your range. When the pass is directly overhead, you can expect the duration of visibility to be twice the range (degrees) divided by the velocity (degrees per minute). For OSCAR 8 this is 16.7 minutes ($2 \times 29.1 \div 3.49$). For the RS birds this duration is 25 minutes ($2 \times 37.7 \div 3.0$). When the pass is halfway between the station and the range limit, the duration is 85 percent of the overhead value; at the range limit, it may be only a minute or less.

A helpful exercise in developing your understanding of how the satellite moves through any portion of your range circle is to place your great circle marker on your globe at various longitudes within the index window and note the path it defines as it comes into and leaves your range circle. For stations in the Northern Hemisphere this track causes the RS birds to depart off to the northeast on ascending orbits and to arrive from the northwest on descending orbits. For OSCAR 8 the opposite relationship results from the difference in inclination angle, its having an angle of inclination 9 degrees greater than 90, while the Russian orbits are 7 degrees less than 90 (see fig. 1). By attaching a length of string to your globe at the station location and tying a knot at a distance representing the range limit, you can obtain an excellent graphical representation of your range limit. Note that the length to the knot can be measured in either latitude degrees or in longitude degrees at the equator.

calculator values

By making the constant for the period and increment that value given earlier, you are simultaneously adding two hours and subtracting a few minutes to the mixed hours and minutes expression. Occasionally, when the minutes value is small, the result comes out a few minutes less than 99 when it should be the same number of minutes less than 59. Subtracting 40 minutes fixes it up.

No doubt a home computer with satellite locating programs can provide more exact information, but satisfactory operation can be achieved using the simple methods described here. For almost ten years I have tracked the satellites — and had hundreds of QSOs — using only my globe, a rubber band, some string, and a hand calculator.

ham radio

working W5LFL from space

All you'll need is ten watts and an antenna — and lots of luck.

So you want to work an **astronaut from space?** You'll have your chance when Dr. Owen Garriott, W5LFL, begins transmitting from the space shuttle *Columbia* on the third day of the nine-day mission scheduled for launch September 30.

Details of the flight plan, are still being worked out. As of this writing, transmission is expected to be within the range of 145.51 to 145.77 MHz, with 145.55 the primary channel (downlink). Using split operation, W5LFL will listen between 144.91 and 145.09 MHz, stepping in 20 kHz increments. (Stay tuned to AMSAT nets and W1AW for specifics. Newsletters will also provide details as the launch date approaches.)

NASA has mapped out a flight path that takes the *Columbia* over the most heavily populated areas of the earth. In ascending orbit, the ship will appear over the southwest horizon of the United States and follow a gentle curve towards the northeast, its path dividing the contiguous United States into four areas corresponding roughly to the time zones. Tracking data is being compiled for release by AMSAT.

Flying at 17,000 miles per hour at an altitude of approximately 200 statute miles, Garriott will have line-of-sight for about 1,000 miles. He'll be radio-visible for a *maximum* of eight minutes over any given point.

Based on experience with OSCAR, AMSAT tells us we can expect to receive full quieting signals while the ship is passing overhead — even on a hand-held radio with a rubber-duckie antenna. Transmission may not be as simple; you'll need ten watts and a gain antenna — and lots of luck — to stand a chance of being herad by W5LFL.

ing heard by W5LFL.

Garriott will operate from the aft flight deck of the *Columbia*, using a specially-built transceiver operating at 5 watts. His split-ring antenna, housed in a window overlooking the European-built Space Lab, will give him line-of-sight perspective to earth during most of the flight.

Astronauts work a schedule of twelve hours on, twelve hours off. Garriott has NASA's OK to operate up to one hour per day during his free time, most likely on either end of his eight-hour sleep; he'll tell Mission Control when he's ready to operate. Club stations at each NASA site will pass the word to other Amateurs; those of us on earth can expect about 35 to 40 minutes' notice.

Once he begins operating, Garriott will transmit on the even minutes and listen on the odd minutes. During transmission, he'll state his location, then identify call areas for which he'll be standing by, and finally begin acknowledging calls he's been able to log.

NASA recommends that amateurs wishing to contact W5LFL transmit *only* their calls during the odd minutes. Just fifteen to twenty orbits at most are expected to be effective, and on each pass, no more than a few dozen earth stations can be worked. If you call but are not acknowledged, don't g.ve up hope: Garriott will tape record all Amateur activity and the resulting tapes will be used as a log for QSLing after completion of the mission. (Attention DX'ers: W5LFL does *not* count as another country!)

September is a good time to clean up that 2-meter equipment. So clean it up, follow the news closely, and get ready to make history . . . as one of the lucky ones to make contact with "W5LFL from Columbia" this fall.

ham radio

by Roy Neal, K6DUE, Senior Correspondent, NBC Network News.

technical forum

Have a technical question? Ask your fellow Amateurs through Technical Forum. You'll reach knowledgeable readers ready to help you find the answer you're looking for. *Have an answer?* Send it in. Every month we'll award a specially selected book from Ham Radio's Bookstore to the writer of the best response to a question previously posed in this column. Address your questions and answers (typewritten, if possible) to Technical Forum, *ham radio magazine*, Greenville, New Hampshire 03048.

efficient matching

Joe Czerniak, W8NWU, asked for "the most efficient way" of feeding a 432 MHz helical antenna with 70 ohm cable from a 50 ohm source (Technical Forum, March, 1983). I assume that by "most efficient way" he meant the way resulting in the least power loss, not the way representing the best possible ratio between personal effort extended and the result achieved.

It seems to me that with the system having a relatively narrow bandwidth, then if effort is not a consideration, a 1/4 wave section of air-spaced coaxial should be constructed to match the unbalanced 140 ohm antenna to the unbalanced line. This should be constructed in the form of a trough, using the jacket of the 70 ohm cable as the outer strips to avoid having a connector (**fig. 1A**). The impedance of the trough section should be 99 ohms (the square root of the product of 140 and 70).

At the equipment end, the matching is best done in the converter, by arranging it to match directly to a 70 ohm feedline.

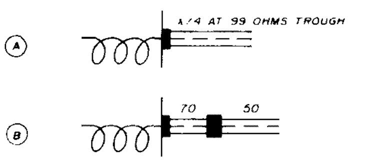


fig. 1. Two methods of matching to a 140-ohm impedance helical: (A) Using a quarter-wave 99 ohm trough, (B) Using successive sections of quarter-wave lines (70 to 50 ohms, respectively).

A more conventional way, using the connectors (while acknowledging their possible future problems), would be to match the antenna to the cable with two successive 1/4-wave sections (**fig. 1B**). It is a fortunate coincidence that a section of 70-ohm coaxial would transform from 140 to 35 (70 is the square root of the product of 140 and 35) and a section of 50 ohm coaxial would transform from 35 to 70 (50 is the square root of the product of 35 and 70). If the helical were a balanced antenna it would be interesting to note that a 140/balanced to 35/unbalanced balun could be built from the same cable used for the feedline.

If the idea of changing the converter input is not acceptable, a pair of nonsynchronous sections, one of 50 ohms and the other of 70 ohms, as described by K1XX in *ham radio* (September, 1978) would be nice. In the same issue, W7VK shows how to put a standard connector on the hardline cable. — **Bob Eldridge, VE7BS**

rain static resolved

HB9FU's letter in July, 1983, Technical Forum recalls difficulty of reception of radio signals aboard aircraft flying through rainstorms. About 1948 an "anti-precipitation static" (APS) wire system began to appear on aircraft.

The wire for the antenna was single-strand copperweld covered with a tough insulation, much like the dielectric in RG-59/U. (In fact, take the outer cover and braid off RG-59/U and you have APS wire.)

While the system on aircraft uses special nylon insulators to grip the wire or to tap off in a "Tee," the main idea is that no rain be allowed to get to the *wire conductor*. Simply stated, the conductor(s) are totally encapsulated.

The static is apparently caused by the charge a wire antenna picks up and the changing charges as the rain drips off.

If HB9FU can get to an airport and see a high-frequency antenna on an overseas aircraft, he'll get the picture — maybe he can also find out where he can obtain that kind of wire and insulators.

Anyway, as long as no part of the conductor is exposed, APS will almost totally disappear. — **Dave Walsh, W1FYX**

LED puzzle

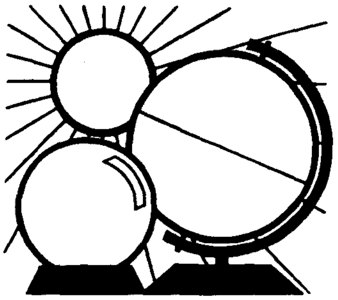
Can anyone tell me what's wrong with the first LED in my Kenwood TS590SE's digital display. It reads *1* on 80 to 40, *2* on 20 and WWV, and *3* on 15 or 10. Can you help? — **John E. Dunn, WB9JKV.**

"hidden" antenna

I would like to correspond with anyone who has used, or is successfully using a "hidden antenna." — **Michael H. Landwehr, KE7T**, P.O. Box 4502 Huachuca City, Arizona 85616.

Heath HX1681

I want to modulate my Heath HX1681 CW transmitter. Is a kit (or circuit) available? — **Robert Miller, W2UWO**



Garth Stonehocker, K0RYW

DX FORECASTER

last-minute forecast

The forecast for September emphasizes the end of summertime DX conditions with its longer daylight hours, increased short-skip openings during the sporadic-E season, and evening thunderstorm QRN. Fall DX conditions are quite different.

In the band-by-band summary below look for geomagnetic-ionospheric disturbances to increase (especially during 1983 and 1984); transequatorial openings to begin again; transpolar openings (gray-line DX) to occur near twilight; maximum usable frequencies to increase; and thunderstorm activity to change from airmass to frontal, with related QRN changes. Accompanying these changes, the lower hf bands (30, 40, 80, and 160 meters) are forecast to be best during the first two weeks of the month. Then the higher bands (30, 20, 15, and 10 meters) are expected to take over, providing excellent conditions considering this point in the 11-year sunspot cycle. Disturbed geomagneticionospheric conditions are probable on September 2-5th, 10-13th, 16-19th, and 23-25th.

The full moon will occur on September 22nd and its perigee on the 6th. The autumnal equinox will be on the 23rd at 1442 UT. No significant meteor showers are expected in September.

solar cycle

It's time to check the decline of this solar cycle. April 1982's lower solar flux and sunspot number (SSN) values confirmed that the gradual downward trend of the cycle from the peak had begun (in February for SSN and April for solar flux). In September 1982 the SSN was recorded at 100.6. At present (June, 1983) a plateau has been reached where it levels off before descending at a more leisurely rate to the minimum years of 1985 and 1986. The plateau is at approximately a SSN of 62 or flux value of 110. It is expected to remain constant through the winter months before decreasing again.

In July I mentioned that the critical frequency, foF2, the main variable of the maximum usable frequency (MUF), was nearly a linear function of SSN or radio flux and gave three representative values at mid-latitude. The foF2 (MUF equals approximately 2.5 × foF2) variation is 0.0485 MHz per SSN value or 0.0557 MHz per flux value, ϕ. The equations are:

$$foF2 = 0.0475\,SSN + 5.25$$
$$foF2 = 0.0557\,\phi + 1.8$$

where *SSN* is the 12 month smoothed sunspot number and ϕ is the monthly mean solar flux. The 2.5 (in the parentheses) is the MUF factor, a geometric angular relationship formed by the distance between the transmitter and the receiver and the height of the ionosphere.

Daily foF2 show greater variation with daily solar flux values or five-day running mean SSN. These relationships will be provided in a future column.

band-by-band summary

Ten and fifteen meters will be open to many areas of the world from morning until early evening hours many days. Geomagnetic disturbances will limit the number of signals heard, but listen carefully—they can be from very unusual places, particularly one-long-hop transequatorial DX. Fifteen meters should stay open later in the day than 10 meters. Operate 10 first and then move down to 15.

Twenty meters will be the main daytime DX band, as it is almost always open to some part of the world. It opens to the east as the sun rises and extends into the late evening hours to the west. Geomagnetic disturbances do not affect this band as much as 10 and 15 meters, but look for unusual transequatorial DX of 5,000 to 7,000 miles (8,000 to 11,200 km). These paths may be possible in the late evening hours during some of these unusual conditions.

Thirty meters is a day and night band. The day portion should be like 20 meters except that signal strengths may decrease during some middays. Days of decreasing strength are related to those having high solar flux values. This band can also be used well into the night, and often through the night; nights in which this doesn't hold true will most likely be those that follow very high solar flux days. The questionable periods are usually the hour or so before dawn. The workable distance may be expected to be greater than 80 meter DX at night and less than 20 during the day.

Forty and eighty meters will exhibit short skip conditions during daylight hours and lengthen after dark. The bands will open for DX to the east just before your sunset, swing more to the south toward Latin America about midnight, and end up in Pacific areas during the hour or so before dawn. On some nights these bands will have low QRN similar to the winter DX season conditions. Coastal regions usually have the edge for working rare DX on these bands. Look for transpolar openings on these bands. (See *ham radio*, September, 1982, p. 56.)

One sixty meters will probably have many nights that will remind you of last summer's noise. However, useful nighttime conditions aren't too far off. Propagation on 160 meters will approximate a shortened 80-meter condition.

ham radio

WESTERN USA

GMT	PDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	5:00	20	20	15	10	15	10	10	15
0100	6:00	20	20	15	10	15	10	10	15
0200	7:00	20	30	20	10	15	10	10	15
0300	8:00	20*	30	20	15*	15	10	10	15
0400	9:00	20	30	20	15	20	10	10	15
0500	10:00	-	20	20	20	20	10	15	20
0600	11:00	-	-	20	20	20	15	15	20
0700	12:00	-	-	20	20	20	15	15	20
0800	1:00	-	-	20	20	20	15	20	20
0900	2:00	-	-	20	20	-	20	20	30
1000	3:00	-	-	-	-	-	20	20	30
1100	4:00	-	-	-	-	-	20	20	-
1200	5:00	-	-	-	-	-	20	20	-
1300	6:00	-	-	-	-	-	20	-	-
1400	7:00	-	-	15	-	-	-	-	-
1500	8:00	-	-	15	20	-	-	-	-
1600	9:00	30	20	15	20	-	-	-	-
1700	10:00	30	15	10	20	-	15	15	-
1800	11:00	30	15	10	15	15	15	15	20
1900	12:00	30	15	10	15	15	15	15	20
2000	1:00	20	20	10	10	15	10	15	20
2100	2:00	20	20	10	10	15	10	10	20
2200	3:00	20	20	15	10	15	10	10	20
2300	4:00	20	20	15	10	15	10	10	20

SEPTEMBER

MID USA

GMT	MDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)	CDT
0000	6:00	30	20	15	10	15	10	10	15	7:00
0100	7:00	30	20	15	10	15	10	10	15	8:00
0200	8:00	30	30	20	10	15	10	10	15	9:00
0300	9:00	30	30	20	10	20	10	10	15	10:00
0400	10:00	30	30	20	15	20	10	15	20	11:00
0500	11:00	30	30	20	15	20	15	15	20	12:00
0600	12:00	30	-	20	15	20	15	15	20	1:00
0700	1:00	-	-	20	20*	20	15	15	20	2:00
0800	2:00	-	-	20	20	20	20	20	-	3:00
0900	3:00	-	-	20	20	-	20	20	-	4:00
1000	4:00	-	-	20	20	-	20	20	-	5:00
1100	5:00	-	-	-	20	-	20	-	-	6:00
1200	6:00	-	-	-	-	-	-	-	-	7:00
1300	7:00	-	-	-	-	-	-	-	-	8:00
1400	8:00	-	-	15	-	-	-	-	-	9:00
1500	9:00	20*	20	15	15	-	-	-	-	10:00
1600	10:00	20	20*	10	15	-	-	-	-	11:00
1700	11:00	20	15	10	15	-	15	15	-	12:00
1800	12:00	20*	15	10	15	15	15	15	-	1:00
1900	1:00	-	15	10	10	15	15	15	-	2:00
2000	2:00	-	15	10	10	15	15*	15	20	3:00
2100	3:00	-	20	10	10	15	10	10	20	4:00
2200	4:00	-	20	15	10	15	10	10	20	5:00
2300	5:00	-	20	15	10	15	10	10	20	6:00

EASTERN USA

GMT	EDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (Caribbean S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	8:00	30	20	15	10	15	10	10	15
0100	9:00	30	20	15	10	15	10	10	15
0200	10:00	30	30	20	10	15	10	10	15
0300	11:00	30	30	20	15*	20	10	10	15
0400	12:00	30	30	20	15*	20	15*	15	20
0500	1:00	30	30	20	15	20	15	15	20
0600	2:00	-	30*	20	15	20	15	15	20
0700	3:00	-	-	20	20	20	15	20	-
0800	4:00	-	-	20	20	-	20	-	-
0900	5:00	-	-	20	20	-	20	-	-
1000	6:00	-	-	-	20	-	-	-	-
1100	7:00	-	-	-	-	-	-	-	-
1200	8:00	-	-	-	-	-	-	-	-
1300	9:00	-	-	-	-	-	-	-	-
1400	10:00	20	-	15	20	-	-	-	-
1500	11:00	20	15	10	20*	-	20	-	-
1600	12:00	20	15	10	15	-	15	-	-
1700	1:00	20	15	10	15	20	15	15	-
1800	2:00	-	15	10	15*	20*	15	15	-
1900	3:00	-	15	10	10	15	15	15	-
2000	4:00	-	15	10	10	15	10	15	20
2100	5:00	-	20	10	10	15	10	10	20
2200	6:00	20	20	15	10	15	10	10	20
2300	7:00	30	20	15	10	15	10	10	20

*Look at next **higher** band for possible openings.

product REVIEW

TH5MK-2 broadband thunderbird

Over the past three years I have been using the Hy-Gain TH5DX triband beam. The performance overall has been excellent with no problems experienced. I had no intention of replacing the antenna — that is until Alan Caplan, W0RIC, of Hy-Gain called and told me about a new version of the TH5 they were getting ready for release. The new antenna is called the TH5MK-2 and is based upon a dual-driven element design. This design facilitates operation from band edge to edge with low SWR and high performance.

For years, broadbanding antennas was not as important as it is now. But today's solid-state radios and amplifiers are much more sensitive to antenna mismatches. To achieve maximum equipment performance, it is imperative to keep VSWR levels below 2 to 1. Otherwise, protective circuits will start shutting the power amplifier down. Another limitation of older antenna designs is that the user usually had to make a choice of either CW or SSB subbands when assembling the antenna. Hy-Gain's solution for broadbanding was the TH5MK-2 — And, its's quite a solution.

construction

My first feeling upon opening the two shipping boxes was that I was looking at a hardware store's inventory. I was overwhelmed by a lot of parts. Before you touch *anything*, grab the instruction manual and read it thoroughly. Hy-Gain has carefully researched instruction manuals and found that when owners read through the instructions several times, assembly problems are drastically reduced. This is a sound recommendation that should be adhered to — without exception. Hy-Gain went to quite a bit of time and expense to write an accurate and precise instruction manual. You'll save yourself time and hassle if you're familiar with the instructions before you start.

Parts are in poly bags so it will also be worth your time to segregate all the different clamps, bolts, nuts, screws, and tubing prior to putting the antenna together. This saved me time and eliminated frantic searching while putting the antenna together. One word of warning. Make sure you carefully measure tubing lengths as you assemble the elements. (I didn't. Well, so much for my advice about reading the manual!)

One place where I did divert (again, not following my own recommendation . . .) from Hy-Gain's instructions was in the assembly of the elements. Hy-Gain suggests that you first build the boom and then assemble the elements on the boom. Not having the time on weekends to sit down and assemble the antenna from start to finish, I decided to assemble the elements first and then mate them to the boom. This allowed me to put the antenna together over several evenings. Other than investing a bit of extra diligence, I didn't have any additional problems using this method. I also didn't have to worry about losing element parts. It took less than two hours to mate all the elements to the boom assembly and assemble the beta match.

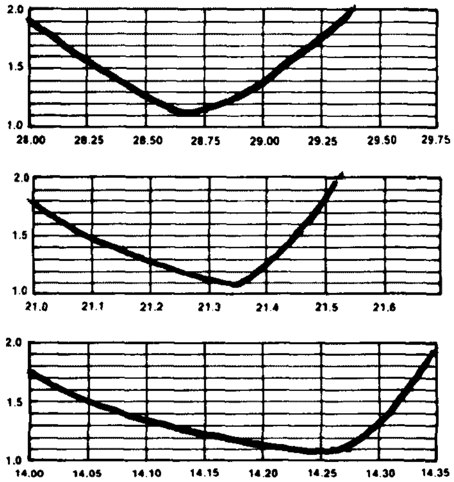


fig. 1. VSWR on 10, 15, and 20 meters (Hy-Gain TH5MK-2).

performance

Initial results have been most favorable with the new TH5MK-2. Not being a "hot and heavy" high band DXer, I can't give a day-by-day summary of its performance. However, in casual DXing and in several contests, it has performed admirably well. In comparing the TH5MK-2 to my TH5DX, my gut feeling is that the TH5MK-2 has a *slight* edge. The big plus, however, is that the MK-2 operates efficiently from band edge to band edge.

specifications

The TH5MK-2 is a five-element broadband antenna designed to operate on 20, 15, and 10 meters. A big plus for the MK-2 is that all hardware is stainless instead of plated steel. This eliminates (or reduces) the problem of hardware oxidization. Hy-Gain also supplies a BN86 balun with the antenna. The TH5MK-2 weighs 57 pounds and has a wind surface area of 7.4 square feet and is designed to survive a 100 mph maximum wind. Each band has one director and one reflector, as well as the two driven elements. The boom is 19 feet long (one foot longer than the TH5DX) and the longest element is 31 feet 6 inches. Hy-Gain rates the antenna at 7.8 dBi gain for 20 meters, 8 dBi for 15 meters, and 9 dBi for 10 meters. Since the antenna employs a Beta match and balun, the antenna is at dc ground for lightning protection. For more information contact Hy-Gain, 9100 Aldrich Avenue, Minneapolis, Minnesota 55420.

N1ACH

Triplett model 3500

The Model 3500 is a compact, portable autorange manual digital multimeter designed to be easy to operate and read and also be able to withstand rough handling. The autorange feature eliminates the need for manual selection of range; you can select the ranges manually if you desire. All measurements except 10 amperes are made from two jacks, eliminating the need to switch test leads as different ranges are selected. The Model 3500 is fully overload-protected with a special 0.3A/250V and 2A/600V fuse arrangement.

The Model 3500 measures ac/dc volts, ac/dc current and high or low power ohms. Accuracy is guaranteed for one year. Protection from overload is up to 1000 Vdc and 750 Vac without fuseblow. The 3500 also features a tilt stand to facilitate greater visibility on the test bench. The case is ribbed to provide a stable, non-slip grip for the user. Battery life is approximately one year in occasional service but less in daily use. The low-battery indicator will alert the user when battery life remaining falls to approximately 50 hours.

We found the Triplett Model 3500 simple to use, handy to have, and rugged enough to stand up well in most ham shacks. Its large LCD display is readable under a variety of lighting conditions. Packaged in a high-impact plastic case, the Model 3500 measures 3-1/2 × 6 × 1-5/8 inches (88 × 155 × 34 mm). It weighs 12.5 ounces (0.35 kg) with battery installed. The suggested retail price ($140) includes the unit itself, 36 inch test leads with alligator clips, and two 1-1/2 volt batteries. Contact your local distributor or Triplett Corp., One Triplett Drive, Bluffton, Ohio 45817 for further information. RS#301

N1ACH

NEW products

1.2 GHz mobile transceiver and receiver

ICOM's new IC-120 is a 1.2 GHz fm mobile transceiver, covering 1260 to 1300 MHz. This unit is styled similarly and has features similar to the IC-15A/H series of 2-meter transceivers, and has many equivalent features. Duplex split is variable, but is initiated at 20 MHz when the unit is first turned on. Duplex up and down as well as scanning features are offered. Power output is 1 watt. ICOM is the first to offer a full-featured mobile transceiver for this mostly unused band.

To complement ICOM's entry into the 1.2 GHz Amateur band with its IC-120 mobile transceiver, ICOM has also announced the release of a 1.2 GHz repeater with a power output of 10 watts, CTCSS capability, IDER, and DTMF control. The RP1210 is synthesized to be complementary to the IC-120 transceiver, and has a duplex split of 20 MHz.

For further information, contact ICOM America, Inc., 2112-116th Ave., NE, Bellevue, Washington 98004. RS#302

programmable linear amplifier

The ETO Alpha 85 features a microprocessor control circuit that eliminates manual tune-up and ensures maximum output and performance. Using the new Eimac 3CX800AF ceramic metal triodes, the Alpha 85 is designed to cover 1.8-2 and 3-22 MHz and is rated at 1500 watts PEP rf output SSB, 1500 watts CW and 1000 watts RTTY and SSTV. (Export and government models tune to 30 MHz. U.S. licensed Amateurs can modify their unit to tune to 29.7 MHz.)

NEW products

The microprocessor is factory-programmed to tune each Amateur band except for 10 meters. Five band-segment pushbuttons can be field programmed to any frequency for rapid band changes and ease of use. The unit can also be manually tuned. To ensure maximum output on all high-frequency bands, a tuned input circuit has been incorporated.

The Alpha 85 measures 7-1/2 × 17-1/10 × 17 inches and weighs 60 pounds.

For more information, contact Ehrhorn Technological Operations, Inc., Canon City, Colorado 81212.

handheld transceiver

The new 2-meter handheld transceiver, model 2591, from Ten-Tec was designed in collaboration with Motorola. Its advanced microprocessor provides features that include a key pad, ten memories, memory scan, and programmable band scan. Any section of band within user-selected upper and lower limits may be scanned in steps of 5, 10, 15, 25, or 30 kHz. Either step size, upper limit or lower limit can be changed independently without complete reprogramming. User selectable HOLD or SKIP is also included. Manual scan is possible — up or down in 5 kHz steps.

A new memory lockout feature allows the scanner to temporarily bypass channels while retaining memory and provides quick lockout of busy frequencies yet returns to normal operation on command.

Other features include selectable power level and extended frequency coverage from 143.500 to 148.995 MHz. A dual function LED shows battery status and indicates transmit mode. A quick-release Ni-Cad pack provides heavy duty 450 mAH rating at 8.4 volts. Standard accessories include rubber flex antenna with BNC connector and wall charger.

The price of Model 2591 is $319.00.

For more information, contact Ten-Tec, Inc., Sevierville, Tennessee 37863.

computer patch™ interface

Now you can convert your personal computer and transceiver into a full-function RTTY station with the new CP-1 Computer Patch™ interface by AEA and appropriate AEA software and cabling. Software packages include split screen operation and large type-ahead and message (brag) buffers at all the common RTTY and CW speeds. No computer programming knowledge is required to use the CP-1.

The CP-1 demodulator is said to provide improved performance compared to single channel RTTY detectors. An easy-to-use AEA

NEW products

magic-eye bargraph tuning indicator gives the closest thing to scope tuning, but separate Mark/Space scope output jacks are also provided. A state-of-the-art multi-usage active filter is incorporated offering pre- and post-limited filtering. Floating comparator (automatic threshold) circuits give the best possible copy under fading and weak signal conditions.

Additionally, the CP-1 offers a variable receiver shift capability for any shift from 100 to 1000 with a NORMAL/REVERSE tone selector switch on the front panel.

A function generator chip is utilized for clean, stable sine wave AFSK tone output to the transmitter. Both plus (+) and minus (−) keyed output jacks are provided for CW keying of virtually any popular transceiver. Automatic transmit/receive switching is available under computer control or from a front panel manual transmit button. Output and computer control signals are available in the usual TTL levels (or RS-232 format with an optional low-cost RS-232 kit).

For complete information, contact Advanced Electronic Applications, Inc., P.O. Box C-2160, Lynnwood, Washington 98036. RS#304

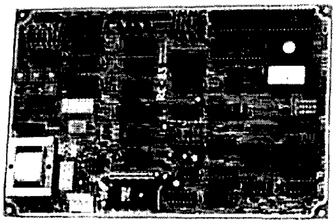
low-cost repeater control

The RC-85 low-cost repeater controller board from Advanced Computer Controls, Inc., is a scaled-down, simplified version of the RC-850 controller. The microcomputer-based control board offers built-in speech synthesis, Touch-Tone remote programming, and a full-featured autopatch at a cost comparable to a collection of "dumb" repeater control boards. It provides the complete interface between the repeater's receiver and transmitter.

The autopatch features autodial, long distance protection, optional phone number readback, DTMF or dial pulse regeneration, and reverse patch.

Some operating parameters of the RC-85 control board may be changed remotely with Touch-Tone commands, including ID and tail messages (with its built-in Touch-Tone activated message editor), command codes, and autodial numbers. Storage is in write-protected RAM. An optional fully documented Personality EPROM allows the repeater owner to define a backup set of parameters independent of the firmware for use in case power is lost. Built-in diode switching allows easy battery backup.

High quality, natural sounding speech synthesis, including a portion of ACC's custom repeater vocabulary, is built in. Together with the control board's Touch-Tone activated message editor, ID and tail messages can be made informative, reminding users of nets, meetings,

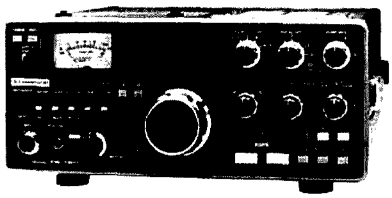


and special events, or informing users of repeater status. An alarm logic input provides the repeater owner with a site alarm to enhance security of the installation. The RC-85 uses CMOS circuitry (no TTL) for low-power consumption from a single + 12 volt supply. All ICs are socketed in high reliability machine contact sockets. The control board is a compact 6″ × 9″ for easy integration into any repeater. The price is $849.

For more information, contact Ed Ingber, WA6AXX, at Advanced Computer Controls, Inc., 10816 Northridge Square, Cupertino, California 95014. RS#305

dual band radio

Trio-Kenwood has recently announced a new model, the TS-780 All-Mode "Dual-Bander," a combination of 2-m/70-cm (144-148 MHz/430-440 MHz) radio designed to meet the needs of the Amateur Radio enthusiast active on both bands. Important features included in the TS-780 are USB, LSB, CW, and FM operation, dual digital VFO's, cross-frequency operating capability, 10 memories that store frequency and band data, internal battery memory back-up, band scan, memory scan, i-f shift, fluorescent tube digital display, 2-m offset switch, and 10 watts of rf output. VOX and



CW semi break-in are built in. An optional TU-4C two frequency tone encoder unit is available.

Additional information is available from Trio-Kenwood Communications, 1111 West Walnut Street, Compton, California 90220. RS#306

flea market

Coming Events
ACTIVITIES
"Places to go..."

CALIFORNIA: Sonoma County Radio Amateurs giant Indoor flea market, Saturday, September 17, 9 AM to 3 PM, Sebastopol Community Center, 390 Morris St., Sebastopol, 5 miles west of Santa Rosa. Free admission and parking. Indoor flea market space $3 ($6 with table) at the door. Advance $2.50 and $5. Vendor set up 8 AM. Radio clinic, refreshments, auction in afternoon. Talk in on 148.13/73. For tickets and information: SCRA, Box 116, Santa Rosa, CA 95404.

CONNECTICUT: The Natchaug ARA will hold a giant flea market, Sunday, September 25, 9 AM to 4 PM, Elks Home off Rt. 32, Willimantic. Indoor/outdoor, rain or shine. Dealers set up 8 AM. Refreshments, free parking. Tables $5 advance, $7 at door. Admission $2.00. Under 16 free. For information: Edward C. Sadeski, KA1HR, 49 Circle Drive, Willimantic, CT 06226, (203) 423-7137; or Clifton Pease, KA1HYW, 268 Main Street, Willimantic, CT 06226, (203) 456-1432 after 4 PM.

COLORADO: The Boulder Amateur Radio Club's "Barcfest", September 25, National Guard Armory, 4750 N. Broadway, Boulder, 9 AM to 3 PM. Admission $3.00 per person or $3.00 per family. Free parking. Refreshments. Talk in on 146.10/70 and 146.52 simplex. For information: Tim Groat, KRØU, 1000 East 10th Avenue, Broomfield, CO 80020. (303) 466-3733.

FLORIDA: The Platinum Coast Amateur Radio Society's 18th annual Hamfest and indoor swap-and-shop, September 10-11, Melbourne Auditorium, Melbourne. Admission $3.00 advance, $4.00 door. Swap tables and tailgating space available. Free parking. Awards, forums and meetings. Talk in on .25/.85 and .52/.52. For reservations, tables and information: PCARS, P.O. Box 1004, Melbourne, FL 32901.

GEORGIA: The 10th annual Lanierland ARC Hamfest, September 25, 9 AM, Holiday Hall, Holiday Inn, Gainesville. Free tables and inside display for dealers. Large flea market, boat anchor auction. Free admission. Talk in on 146.07/.67. For information: Phil Loveless, KC4UC, 3574 Thompson Bend, Gainesville, GA 30506. (404) 532-9160.

IOWA: The 9th annual Cedar Valley ARC Hamfest, Sunday, October 2. Doors open 7 AM, Hawkeye Downs Exhibition Building, Cedar Rapids. Tickets $2.00 advance, $3.00 door. First table $5.00 others $7.00. Overnight camping area, picnic facilities, airport nearby. Collins surplus store will be open 9 AM to 2 PM. Talk in on 146.16-.76, .52, 223.34-.94. For tickets and information: CVARC Hamfest, P.O. Box 994, Cedar Rapids, IA 52406.

ILLINOIS: The Sangamon County Fair and Agricultural Association's 6th annual New Berlin Hamfest, Sunday, September 25, Sangamon County Fairgrounds, New Berlin, rain or shine. Camping available. A model railroad train meet will also be held that day. For details: K9QFR, Box 2, Pleasant Plains, IL 62677.

ILLINOIS: The Peoria Area ARC Superfest '83, September 17 and 18, Exposition Gardens, W. Northmoor Road, Peoria. Gate opens 6 AM. Commercial building 9 AM. Admission $3.00 advance, $4.00 gate. Amateur Radio and computer displays, huge free flea market, free transportation to Northwoods Mall on Sunday. Camping facilities on grounds. Talk in on 146.16/76, W9UVI. For information and reservations SASE to: Superfest '83, 5808 N. Andover Ct., Peoria, IL 61615.

KANSAS: The Sandhills ARC's annual Eye-Ball QSO Party, September 25, Finney County Fairgrounds, Garden City. Doors open 9 AM. For information SASE to S.H.A.R.C., P.O. Box 811, Garden City, KS 67846.

INDIANA: The 4th annual Grant County ARC Hamfest, Saturday, Septmeber 10, McCarthy Hall, St. Paul's Catholic Church, Marion. Doors open 8 AM. Donation $2.00 advance, $3.00 gate. Table reservations $2.00 per 8 ft. table. Refreshments, free parking. Talk in on 146.19/79 or 146.52 simplex. For information/tickets SASE to: KA9DLJ, Jerry Richards, P.O. Box 1146, Marion, IN 46952.

MARYLAND: The Columbia Amateur Radio Association's 7th annual Hamfest, Sunday, October 23, 8 AM to 3:30 PM, Howard County Fairgrounds, 15 miles west of Baltimore. Admission $3.00. Indoor tailgating and tables $6.00 additional. Outdoor tailgating $3.00 additional. Talk in on 147.735/135, 146.52/52. For reservations and information: Ed Wallace, K3EF, 9905 Carillon Drive, Ellicott City, MD 21043.

MASSACHUSETTS: The 19-79 ARA's annual fall flea market, Sunday, October 16, 11 AM to 4 PM, Beachmont VFW Post, 150 Bennington St., Revere. Sellers 10 AM. Admission $1.00. Sellers tables $6.00 advance, $8.00 at door if available. Talk in on 19-79 and 52 direct. For table reservations send check to 19-79 Amateur Radio Association, P.O. Box 171, Chelsea, MA 02150.

MICHIGAN: The Grand Rapids ARA's annual Swap and Shop, Saturday, September 17, Hudsonville Fairgrounds. Dealers. Refreshments. Indoor sales area and outdoor trunk swap area. Gates open 8 AM. Talk in on 146.16/76. For information: Grand Rapids ARA, P.O. Box 1248, Grand Rapids, MI 49501.

MICHIGAN: The L'Anse Creuse Amateur Radio Club's 11th annual Swap and Shop, Sunday, September 18, 9 AM to 3 PM, L'Anse Creuse High School on Reimold. Admission $1.00 advance, $2.00 at door. Good food, free parking. Talk in on 147.69/.09 and 146.52. For information and tables SASE to William Chesney, N8CVC, 215 Elizabeth @ Mt. Clemens, MI 48043. (313) 463-1412.

NEW ENGLAND: The Hosstraders will hold their annual autumn swapfest, Saturday, October 8, rain or shine, Deerfield, New Hampshire, Fairgrounds. Admission $1.00, tailgating included. Friday night camping for self-contained rigs, after 4 PM. Profits benefit Shriners' Boston Burns Hospital. Last May's donation $2,702.00. For information/map SASE to Norm, WA1IVB, RFD Box 57, West Baldwin, ME 04091; or Joe, K1RQG, Star Rt. Box 56, Bucksport, ME 04416; or Bob, W1GWU, 105 Walton Rd., Seabrook, NH 03874.

NEW JERSEY: The De Vry Technical Institute ARC's annual flea market, October 1, 9 AM to 4 PM, school parking lot at 479 Green Street, Woodbridge. Buyers free admission. Sellers $3.00. No electricity. For information *contact* WB2JKU.

NEW HAMPSHIRE: The Connecticut Valley FM Association's 6th annual Hamfest and Flea Market, September 25, 9 AM to 5 PM, King Ridge Ski Area, Sutton, rain or shine. Admission $2.00. Dealers and flea marketeers $5.00. Refreshments available. Overnight camping for self-contained RV's. No hookups. Talk in on 146.16/76 or 146.52 simplex.

NEW YORK: The Yonkers electronics fair and giant flea market, Sunday, October 2, 9 AM to 4 PM, Yonkers Municipal Parking Garage, Corner of Nepperhan Avenue and New Main Street. Rain or shine. Live demonstrations, Amateur Radio, computers, electric car, satellite TV, SSTV, Hi-Fi/audio. Refreshments, free parking. Unlimited free coffee all day. Admission $2.00, children under 12 free. Sellers $6.00 per space, bring tables. Sponsored by the Yonkers ARC. For information (914) 969-1053.

NEW YORK: The Radio Amateurs of Greater Syracuse, RAGS, annual Hamfest and Computer Display, Saturday, October 1, 9 AM to 6 PM, Art and Home Center, New York State Fairgrounds, Syracuse. Commercial exhibitors, large indoor/outdoor flea market, tech talks, ARRL booth, contests and entertainment. Refreshments. Admission $3.00. Talk in on 90/30, 31/91 and 52 simplex. For information: RAGS, Box 88, Liverpool, NY 13088.

NEW YORK: The Elmira International Hamfest, September 24, Chemung County Fairgrounds, gates open 6 AM. Free flea market, dealers, tech talks. Refreshments available. Talk in on 147.96/36, 146.10/70 and 146.52/52. Advance tickets $2.00 from John Breese, 340 West Avenue, Horseheads, NY 14845. Tickets at gate $3.00.

NEW YORK: Seaway Valley Hamfest, Saturday, September 10, 9 AM to 4 PM, Louisville Firemen's Arena, Louisville. ARRL forum, tech programs, demos, entertainment, flea market, snack bar. Tickets $3.00 at door, $2.50 advance. For information/tickets: SASE to Lois Ierlan, WA2RXO, 725 Proctor Avenue, Ogdensburg, NY 13669. (U.S. funds only please).

NEW JERSEY: The South Jersey Radio Association's 35th annual Hamfest, September 18, Pennsauken Sr. High School, Hylton Road, Pennsauken, 8 AM to 4 PM. Advance tickets $2.50. $3.50 at gate. Tailgaters $5.00. Refreshments available. Talk in 22/82 and 52. For information: Fred Holler, W2EKB, 348 Bortons Mill Road, Cherry Hill, NJ 08002. (609) 795-0577.

OHIO: The Cleveland Hamfest and ARRL Great Lakes Division Convention, Saturday, September 24, banquet. Sunday, September 25, Hamfest, 8 AM to 5 PM. NEW LOCATION: Cleveland Aviation High School, North Marginal Road, Cleveland. Exhibits, refreshments, overnight parking available. Flea market open 6 AM, $2.00 per space. General admission $3.00. Talk in on 146.52, W8QV. For advance tickets send check or MO for $2.50 before August 31 to: Cleveland Hamfest Association, P.O. Box 93077, Cleveland, OH 44101.

WASHINGTON: The Walla Walla Valley Radio Amateur Club's 37th annual Hamfest, September 24 and 25, Milton-Freewater, Oregon Community Building. Doors open 8:30 AM. Swap shop, crafts, Tesla coil and spark gap demo. Antique, homebrew contests. Bazaar Sunday. Potluck dinner Sunday noon. For information: W7DP, P.O. Box 321, Walla Walla, WA 99362.

PENNSYLVANIA: 1983 A-5 Magazine/USATVS 4th annual fall ATV Conference, September 23-25, to be held concurrent with York, PA Hamfest. Seminars Friday 6-10:30 PM. Saturday 9 AM to 10 PM with banquet. Sunday 7 AM to noon. FSTV and SSTV to be fully covered with excellent speakers. SASE to W3SST or ATV Magazine, P.O. Box H, Lowden, IA 52255.

PENNSYLVANIA: The Skyview Radio Society's annual Hamfest, Sunday, September 18, noon to 4 PM, club grounds on Turkey Ridge Road, New Kensington. Talk in on 04-64 and 52 simplex. Registration fee $2.00. Vendors $4.00.

PENNSYLVANIA: The Pack Rats (Mt. Airy VHF ARC) invites all Amateurs and friends to the 7th annual Mid-Atlantic VHF Conference, Saturday, October 1, Warrington Motor Lodge, Route 611, Warrington; and Sunday, October 2, Hamarama, Bucks County Drive-In Theater, Route 611, Warrington. Flea market admission $3.00. Selling space $5.00 each. Bring own tables. Gates open 7:30 AM. Rain or shine. Advance registration $4.00 for conference and Hamarama. Send to Hamarama '83, P.O. Box 311, Southampton, PA 18966 or Lee A. Cohen, K3MXM, (215) 635-4942.

SOUTH CAROLINA: 32nd annual Rock Hill Hamfest, October 2. For information: YCARS, Box 4141CRS, Rock Hill, SC 29730.

SOUTH DAKOTA: The ARRL Dakota Division Convention, September 23-25, Howard Johnson Motor Lodge, Sioux Falls. Friday evening registration and entertainment. Saturday forums, exhibits, flea market, contests, luncheon. ARRL forum lead by Vic Clark, President. Sunday 2 meter transmitter hunt. Convention pre-registration prior to September 1 is $5.00. $6.00 at door. Convention and banquet $15, $16 at door. Banquet only $10. Talk in on 16/76 and 52/52. For information, Sioux Falls ARC, P.O. Box 91, Sioux Falls, SD 57101.

RADIO EXPO: Sponsored by the Chicago FM Club, Saturday and Sunday, September 24 and 25, Lake County Fairgrounds, Routes 120 and 45, Grayslake, Illinois. Flea market opens 6 AM. Exhibits open 9 AM. Indoor flea market tables available at $5.00 per day. Tickets $3.00 advance, $4.00 at gate, good for both days. Seminars, tech talks, ladies' programs. Talk in on 146.16/76, 146.52 and 222.5/224.10. For information: SASE to Radio Expo 83, Box 1532, Evanston, IL 60204 or (312) 582-6923.

TENNESSEE: The Memphis Hamfest, Saturday and Sunday, October 8 and 9, Memphis Fairgrounds, Mid South Building. Radio and computer displays and forums. Hospitality party Saturday night. Dealer and flea market setup Friday evening til 9 PM. Tables on site. Talk in on 28/88 and 34/94. For information/reservations: Clayton Elam, K4FZJ, 28 No. Cooper, Memphis, TN 38104. (901) 274-4418 days or (901) 743-6714 evenings.

TEXAS: The Wichita ARS's second annual Hamfest, September 24 and 25, National Guard Armory, Wichita Falls. Saturday 8 AM to 6 PM. Sunday 8 AM to 2 PM. Dealer displays, computers, large inside flea market, refreshments, free RV parking no hookups. Nearby shopping malls. Air Force MARS, QCWA meeting and more. Pre-registration by September 21, $4.00. $5.00 at door. Talk in on 146.34/94 and 147.75/15. For information/tickets: WARS HAMFEST, P.O. BOX 4363, Wichita Falls, TX 76308.

VIRGINIA: The 8th annual Tidewater Amateur Radio Hamfest, Computer Convention, Electronic Flea Market, Saturday and Sunday, October 8 and 9, Virginia Beach Pavillion, 9 AM to 5 PM. $4.00 admission good for both days. Flea market tables $5.00 one day, $8.00 both days. Dealers, displays, forums, computers, satellite equipment. Beautiful Virginia Beach nearby. Visit Norfolk Waterfront Festival Marketplace. For information/tickets: Jim Harrison, N4NV, 1234 Little Bay, Norfolk, VA 23503. (804) 587-1695.

WISCONSIN: The St. Croix Valley Repeater Association's 7th annual Hamfest, Saturday, September 24, 9 AM to 2 PM, American Legion Hall, Baldwin. Admission $2.00. Talk in on 147.93/.33 and 146.52 simplex. For information: Bruce Olson, N9BLU, Box 91, St. Croix Falls, WI 54024.

OHIO: Lima Hamfest, Allen County Fairgrounds, October 9. Gates open 6 AM. Advanced tickets $3.00; at gate $3.50. Tables $6.00 each; half tables $3.50. For information/reservations: N.O.A.R.C., Box 211, Lima, Ohio 45802.

OPERATING EVENTS
"Things to do..."

SEPTEMBER 3 AND 4: The Independence FM Radio Association will operate a special event station celebrating Santa Caligon Days in Independence. Frequencies: 10-30 kHz from lower General and Novice band edges. Certificate for large SASE to KD0FW, Mike Bogard, 608 Concord Circle, Independence, MO 64056.

SEPTEMBER 13-17: The Southern Counties ARA will operate a special events station at the Miss America Pageant, Caesars Hotel Casino on the boardwalk in Atlantic City on a 24-hour basis. Frequencies: Phone - 25 kHz inside General band. CW 65 kHz from bottom of General band. 80-10 meters. Novice contacts 15 and 40 meters in the middle of Novice band. Local SCARA repeater K2BR at 146.745, output/input down 600. A special QSL available for SASE to: SCARA, K2BR, Box 121, Lynwood, NJ 08221.

SEPTEMBER 17: The Sweetwater ARC will operate special event station WA7USI from an historic site in southwestern Wyoming from 1800Z Sept. 17 to 1800Z Sept. 18. Frequencies: Up 40 kHz on General phone bands. Also Novice.

SEPTEMBER 17-19: Kansas State QSO Party sponsored by Boeing Employees' ARS of Wichita. Sept. 17 - 0100 UTC to 0700 UTC. Sept. 17 1300 UTC to 0700 UTC Sept. 18. Sept. 18 1300 UTC to 0100 UTC Sept. 19. All Amateurs are invited to participate. All bands and modes may be used. Kansas stations send QSO number, RS(T) and county. All others send QSO number, RS(T) and state, Province or foreign country. Logs must show dates, times in UTC, stations worked, exchanges, bands, modes and scores claimed. Log and summary sheets available for SASE. Must be postmarked NLT October 20, 1983 and sent to: Boeing Employees' ARS of Wichita, c/o Mike Thornton, WA0TAH, 1001 Munnell Avenue, Wichita, KS 67213.

SEPTEMBER 18: The Wisconsin Valley Radio Association will operate a special event station from Marathon County in north central Wisconsin at the intersection of 45 degrees North Parallel and 90 degrees West Meridian, near the city of Wausau, exactly halfway between the North Pole and the Equator and halfway between the Zero Meridian at Greenwich, England, and the International Dateline. Listen for club call W9SM. 7 AM to 7 PM CDT. Frequency (depending on band conditions) will be 25 kHz up from bottom of General phone portion of band(s) used. For a QSL card SASE with $1.00 to Wisconsin Valley Radio Association, Box 363, Wausau, WI 54401.

SEPTEMBER 23 AND 24: The Smithfield (Ohio) Apple Festival is sponsoring a special event station. 2300 UTC to 0400 UTC. Frequencies: SSB, 3.900 ± 5 MHz. Novice 7.110 ± 5 MHz. Station call N8CUX. For a special certificate SASE to Robert Carson, N8CUX, 259 Hill Street, Smithfield, Ohio 43948.

SEPTEMBER 22-24: The Fisher Body Lordstown ARC will operate W8KKZ from 1200Z to 0300Z daily to commemorate the 75th anniversary of General Motors. Operations lower portion of 20 and 40 meter General phone band and lower portion of 15 and 40 meter Novice CW bands. QSL information will be given on the air.

OCTOBER 1 AND 2: Oregon QSO party sponsored by the Hermiston Amateur Radio Club. Oct. 1 1700Z to 0800Z Oct. 2. 1500Z Oct. 2 to 0000Z Oct. 3. Exchange: OR stations signal report and county. Others signal report and state/province/country. A station may be worked once per band and mode. Mixed mode or CW only. All entries must have log and summary sheet. Summary sheet available from KA7IXH for SASE. Logs must be received by November 4, 1983. Mail to Bob Franklin, KA7IXH, Rt. 3, Box 3783, Hermiston, OR 97838.

OCTOBER 1-9: Special events station KN5D will operate throughout the 12th annual International Hot Air Balloon Fiesta, Albuquerque, New Mexico. 1400 to 1800 UTC daily. Frequencies: 15-25 kHz above low end of General phone bands. For a special QSL card and Balloon Fiesta certificate SASE to KN5D, P.O. Box 997, Corrales, NM 87048.

OCTOBER 1,2 & 8,9: VK/ZL/Oceania Contest. The WIA and NZART invite worldwide participation in this year's VK/ZL DX contest. Phone: From 1000 GMT, Saturday, October 1 to 1000 GMT, Sunday, October 2. CW: From 1000 GMT, Saturday, October 8 to 1000 GMT, Sunday, October 9. Logs overseas stations: Show date, time in GMT, call sign of station contacted, band, serial number send, serial number received. Underline each new VK/ZL contact. Separate logs must be submitted for each band. Summary sheet to show call sign, name and address, equipment used, QSO points for each band, VK/ZL call areas worked on that band. WIA VK/ZL Contest Manager, VK3BGW, 1 Noorabil Court, Greensborough, Victoria 3088, Australia by January 31, 1984.

OCTOBER 1983 / $2.50

ham radio magazine

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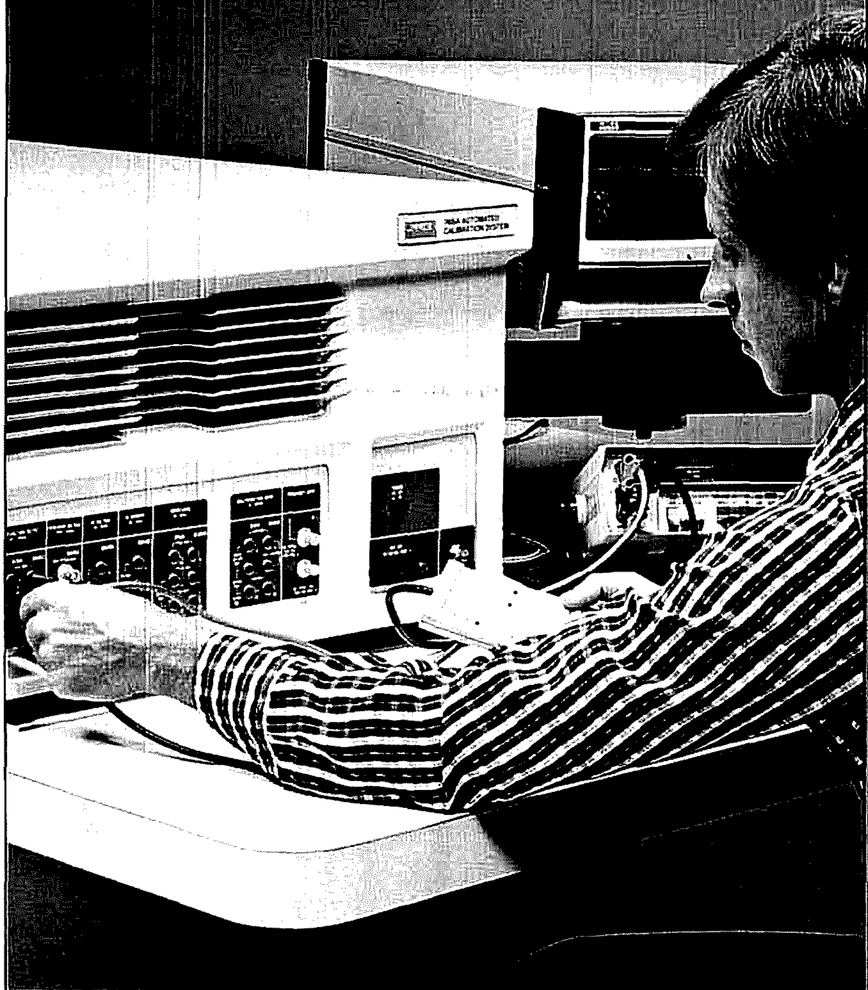
BACK TO BASICS: DEFINING AMPS, VOLTS, AND OHMS



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ham radio

magazine

OCTOBER 1983

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contents

REFLECTIONS

"The Old Man" lives

I don't know how many of you remember "T.O.M." I'm not old enough myself to have read the column when it first came out (1916-1932?) but I sure became another avid fan of his while pouring through old volumes of *QST*. For those of you who have never had the pleasure, let me introduce you to one of the greatest cynical voices of Amateur Radio to have ever existed.

"T.O.M." was a pseudonym assumed by an individual who was gutsy enough to call it as he saw it. He recognized situations for what they were and freely voiced his opinion. Usually the subjects were associated with either poor operating techniques or poorly functioning equipment. The comments in the form of a letter to the magazine would mysteriously appear on the editor's desk without any knowledge of who wrote it or where it had come from. There are those who believe it was written by Hiram Percy Maxim himself. In honor of the fine gentleman, whoever he may have been, I dedicate the following short piece and actively solicit comments.

Rotten operating. At the top of my list are LISTS. Hate is a strong word, but it sure approaches how I feel about self-appointed "gods" who believe it is their inalienable right to "manage" a DX station's callers when they themselves have a hard time understanding the DX station because of low signal level *at their* location or a language "barrier." A case in point is a well known French-speaking DX station on 20 meters who puts an excellent signal into the East Coast at least. One of the MC's in this case had a hard time hearing and understanding the station while others were copying him Q5. What sense does it make to take a list of 100 calls when the band is dropping out? What sense does it make to have two or three MC's trading off calls on the *same* frequency while the DX station just waits? I'm not talking just seconds — I'm talking many minutes — precious minutes as the vagaries of propagation does its thing.

In defense of list operation there are those who say that the DX station requested help and couldn't handle the ensuing pile-up alone. I used to feel that way. But come on, folks! What is the real value of that contact if your call has been *re*transmitted by a more powerful station and perhaps even your signal report? It cheapens the process and product. (Incidentally, what is the product? I won't get into that now. I'm not done with list operation yet.)

Dear "rare" DX station: If you have a hard time sorting out the horde and want to use an MC, why don't you tell him to repeat the following: "The DX station requests all callers to listen carefully, keep all calls short and only answer if your letters are acknowledged." Listen to Arild, 3X4EX, some day and see how he handles pile-ups — extremely effectively, I might add. If the DX station must have an MC, how about interspersing the list operation with "free" periods during which anyone NOT on the list may call in? The DX station just might be very surprised to see how many more stations he can work and also gain some confidence in the process. By the way, DX station, if there is some persistent Amateur calling out of turn and you just want to get him off your back, acknowledge him and just forget to log the contact, or more directly, *say so at the time*. (That'll either straighten everyone up or create pandemonium.)

I have to admit that on the lower bands there's more of a temptation to assist a friend or "twenty friends." (I've done it myself and have been helped also.) But I've tried to do it right in as rapid fire a manner as possible, not retransmitting calls, and I have been extremely sensitive to even one dissenting voice on the frequency.

There, I've said it. There are many other areas of rotten operating that I can and will address as the need arises. Remember folks — this is our hobby — with a stress on the word *hobby*. Let's go back to basics, recalling that it's a privilege to use the frequencies and try to communicate a little more.

Rich Rosen, K2RR
Editor-in-Chief

P.S. You still have an opportunity to express *your* opinions. Just grab your copy of the September issue, find the survey on the editorial page and go for it. Preliminary returns have been fantastic. I know it's impossible, and too much to wish for, but if everyone could just take the time to fill out the survey what a well-rounded poll that would make. It would help us provide to you the best ham radio magazine of the 80's.

presstop de W9JUV

AMATEUR RADIO-CABLE TV CONFRONTATION CONTINUES TO HEAT UP, with the ARRL filing a Motion for Expedited Action on its previous petition to have cable channels E (2 meters) and K (220 MHz) taken away from the cablecasters. That earlier petition, RM-4040, was filed in January, 1982, and cited numerous problems Amateurs throughout the country were having with cable system leakage. Attempts to resolve the problem through discussions with the National Cable TV Association have met with frustration, hence the latest League action to push the Commission toward a workable solution. The NCTA received an extension of time to file comments on the League's petition, so its response wasn't available at press time.

Excessive Leakage Will Cost A California CATV Firm a $3000 fine, according to the FCC. Times-Mirror Cable Television was cited for leakage problems with its Encinitas (near San Diego) system, which was found to have high signal leakage on many frequencies including the Amateur bands. That system had been investigated previously by the FCC, but had claimed to have cleared up its leakage problems.

OCTOBER 28 IS THE NEW DATE FOR STS-9'S LAUNCH with W5LFL/Space/Mobile aboard, after delays in checking out some of the mission systems. W5LFL's hand-held is a custom-built Motorola, which passed NASA's rigorous compatibility checks after a minor interference problem with a telemetry system was found to be due to cable routing.

W5LFL's Primary Transmit Frequency Will Be 145.550 MHz, with 145.250, 145.530 and 145.570 as back-ups. He'll listen on 144.650, 144.700-144.850 (25-kHz channels), 144.910-145.090 (20-kHz channels), plus 145.350 and 145.450. The 25-kHz channel spacing is to conform with 2-meter usage in other parts of the world. In all cases he should specify the frequencies to which he'll be listening, as well as any special instructions such as call areas or states he's looking for, during his "even minute" transmit period.

EXTENSION OF AMATEUR LICENSE TERMS TO 10 YEARS should be the first Commission action for Amateur Radio following the August recess. Present licenses would not be extended, but all new licenses and renewals will be for the longer period. It's also expected that the present five-year grace period for renewal after expiration will be trimmed to two years.

Some Other Pending FCC Actions May Be Delayed, however, since the Commission has been reduced by law to five commissioners and only four are presently serving, following Anne Jones' departure. Few controversial issues are likely to be resolved until another Commissioner is on board, to avoid the possible stalemate of a two-two split.

OSCAR 10'S MODE B TRANSPONDER WORKS BEAUTIFULLY following turn on August 6. Users from all corners of the world appeared in the first few days, leading to comments that the downlink sounded more like 20 meters than 2—complete with pileups! Signals have improved steadily, and with the final position tweaking at the end of August, it's possible to work through OSCAR 10 well, with an ERP well under 100 watts.

Some Excessively Strong Signals Have Been Noted, as much as 20 to 30 dB above the beacon. Such signals cut down transponder gain, knocking weaker signals out completely. Users are asked to limit power so their downlink signals are no stronger than the beacon.

ARRL CW Bulletins Are Now Carried On OSCAR 10, 145.840 downlink. Both the AMSAT Tuesday night net and Westlink have also been relayed by the new bird.

Mode L Tests Are Set For Early September; listen for the 70 cm downlink signals Thursdays and Saturdays (GMT) ± one hour from OSCAR 10's orbital apogee.

AMSAT's 1983 Annual Meeting Will Be November 12 near Laurel, Maryland. An excellent space symposium is planned; call AMSAT at (301) 589-6062 right away if you want to attend.

Two Russian Satellites, RS-4 and RS-5, are no longer being heard and appear dead.

ARRL WILL ADMINISTER "NO-CODE" LICENSE EXAMS if the FCC adopts no-code and the ARRL becomes the exam program administrator. Responding to questioners at the Pacific Division Convention, League President W4KFC also stated that membership status and privileges of no-code Amateurs would be decided by the League membership.

BEACON POWER HAS BEEN INCREASED TO 100 WATTS OUT in an addendum to the FCC's recent revision of the rules on Amateur power. Beacons had been limited to 100 watts input.

BELGIAN AMATEURS FACE LOSS OF MANY MICROWAVE BANDS if a bill proposed for the Belgian Parliament becomes law. According to G3WDG, 432-434 MHz plus the 23, 13, 9 and 6 cm bands are all involved. ON6AT/RTT solicits messages of support from other Amateurs.

LOG-KEEPING IS NO LONGER REQUIRED OF CANADIAN AMATEURS, after their Department of Communications deleted that part of its rules in a surprise move. However, a log showing CW activity will still be required in applying for some license endorsements.

NO MORE REPEATERS WILL BE COORDINATED ON TO 2 METERS in southern California. Following the lead of the Northern Amateur Relay Council, which recently ended its attempts to coordinate 2 meters due to overcrowding, the 2-Meter Area Spectrum Management Association (TASMA) agreed that the band could accommodate only a few limited coverage repeaters in remote areas. Any uncoordinated machines that appear could be subject to the sanctions against intentional interference cited by former FCC Private Bureau Chief Jim McKinney.

comments

fan letter

Dear HR:

In regard to the heatsink cooling fan suggested by Ed. Marriner, (*ham radio*, Ham Notes, July, 1983), perhaps a better scheme to minimize losses and unnecessary heating would be to substitute a 1- or 2-μF capacitor for the dropping resistor.

This provides a use for the old "bathtub" capacitors (600V) which otherwise seem to end up in the trash barrel.

Fred J. Norvik, W2GH
Albany, New York

repeater etiquette

Dear HR:

The spectacular growth in the number of repeaters has resulted in operating practices we can all be proud of — as usual, the self-discipline of Amateur Radio operators has made itself evident. (A few minutes spent listening to the radio services that do not have stringent licensing standards will convince anyone that Amateurs are doing a good job.)

Some aspects of repeater use could use improvement; consider, for example, the following:

Overuse and underuse of call signs. Once every ten minutes and at the beginning and end of a QSO does the trick. Some Amateurs are beginning to use call abbreviations — typically, omitting the prefix and using only the body of their call. This is certainly unlawful and confusing to the listener.

Non-substantial communications — some repeaters are monopolized by Amateurs who talk by the hour and say very little. They usually break politely when asked to by a "call please" request but return to their meaningless gabbing as soon as the interruption has cleared. Another problem occurs when people continuously update each other on their whereabouts without apparent reason. In one case overheard recently, this type of never-ending QSO involved a man and wife who continually embellished their conversation with expressions of endearment.

The foregoing applies doubly to phone patch conversations. This is an important privilege; it should be guarded carefully.

It should be apparent that we are judged by what we say. I have two non-Amateur associates who ride to and from work with me. Their comments on conversations we overhear suggest what the public thinks. If we expect the public's support in matters of tvi, rfi, and antenna erection, we must at least convince them that we use our privileges meaningfully. There are hundreds of acceptable subjects for discussion, both related to and not related to Amateur Radio; before you push the microphone button, ask yourself if you really have something to say!

Inconsiderate operation — how many times have you heard a station break into a QSO and take over, ignoring one or more of the original participants? I was testing an antenna not long ago with the help of another very cooperative station. Without any warning, a friend of the other operator broke in and interrupted the tests for several minutes. He signed off as if I were not even there. Eventually the tests were completed. Remember, when you are impolite, a lot of people are listening and forming their opinion of you (and Amateur Radio).

There are those who talk only to a fixed group and make the newcomer feel he did something wrong by just trying to say "hello." Single-purpose repeaters and nets and closed repeaters may have to be accepted. However, those who use repeater time and exclude others for no good reason ought to think twice: none of us has the exclusive right to any Amateur frequency.

On the whole, in my opinion, repeater operation deserves a very high mark. With a little more effort, it could get itself into the "straight A" class!

George A. Wilson, Jr., W1OLP
Walpole, Massachusetts

high spirits

Dear HR:

I don't know when I ever enjoyed *ham radio* as much as I did the May, 1983 issue on antennas. I'm in the process of mounting a vertical atop a 70-foot tower; I don't know how well it will work, but I'll try.

The article on "Log-Yagis Simplified" was really great; it was a big help to me. I just had to write and let you know about it.

Keep the good articles coming. I'll be looking for one soon on vertical antennas mounted above the beams on a tower.

Fred Jones, WA4SWF
Louisa, Kentucky

free QSL's

Dear HR:

Help! My poor secretary is being buried in a pile of coupons from *ham radio* readers who've requested the free QSL cards offered in our August advertisement.

We're making every effort to fill all requests as quickly as possible. But because of the tremendous response to our offer, some hams will have to wait.

If you've requested QSL cards but not yet received them, please be patient; we'll mail yours as soon as the next print order is received.

Thanks.

Tom Bluesteen
Advertising Manager
RCA Government
Communications Systems
Camden, New Jersey 08102

Back to basics:
the fundamentals
of measurement

electrical calibration standards

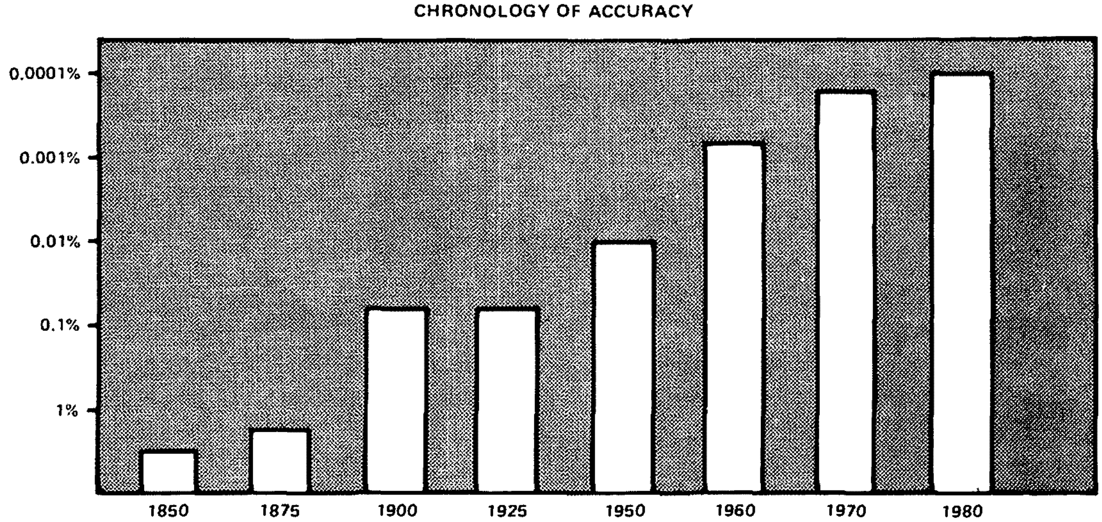

fig. 1. Traceability to NBS (National Bureau of Standards) means the stated accuracy of an instrument has been established through calibration with equipment whose accuracies have been established directly or indirectly by NBS certified references.

Courtesy of the Copyright Owner, John Fluke Manufacturing Co., Inc.

Calibration — whether it be checking the butcher's scales or fine tuning in a radio station — always involves certain principles and goals. These are to achieve accuracy and precision and to permit interchangeability in mass-produced items. Along with metrology, the science of measurement, calibration gives us confidence in the units of measure we use daily.

Before delving into the specifics of determining the five basic electrical quantities (the farad, ampere, volt, henry, and ohm), let's discuss and define some calibration philosophies. Calibration is the process of comparing the readings obtained on test and measurement instruments to those of carefully defined references or standards of usually far greater accuracy in order to determine and then correct any deviations in the instrument. Two benefits result — uncertainty about the particular instrument under test is minimized and the instrument meets the requirements for traceability.

"Traceability" means that measurements can be traced back to the National Measurement System. (See **figs. 1** and **2**.). It exists on seven levels, all traceable to the NBS (National Bureau of Standards):

By Vaughn D. Martin, 114 Lost Meadows, Cibolo, Texas 78108

The Tertiary Level includes instruments used for production work, quality control, maintenance, and general measurement purposes.

Local Secondary Standards are used for calibration work and as references in certain production and engineering areas.

Transfer and Inter-laboratory Standards maintain or represent the basic electrical units locally and are sometimes referred to as Local Primary Standards.

NBS Working Standards are used to calibrate and certify local standards.

NBS Secondary Standards include Transfer, Check, and Scaling Standards. They are used to compare NBS Working Standards to the Legal Electrical Reference Units, the values of which are embodied in the NBS Primary Standards.

NBS Primary Standards are derived from the basic "SI" Units.

The International System of Units (SI) is the "glue" that holds the whole structure together.

what is a standard?

It is possible to build a rugged platinum and iridium standard, but that standard would change, over time, as it became affected by its environment. While a standard could be arbitrarily defined without regard to the physical environment, such a standard would never be more than arbitrary. We could say that it was a standard *derived or based upon a* previous standard, or carefully examine the physical properties and elements of nature and advantageously base the standards on these properties. In actuality, the determination and derivation of each of the various standards employs some or all four of these techniques.

Consider this. The two most basic units of measurement, length and time, are derived as follows:

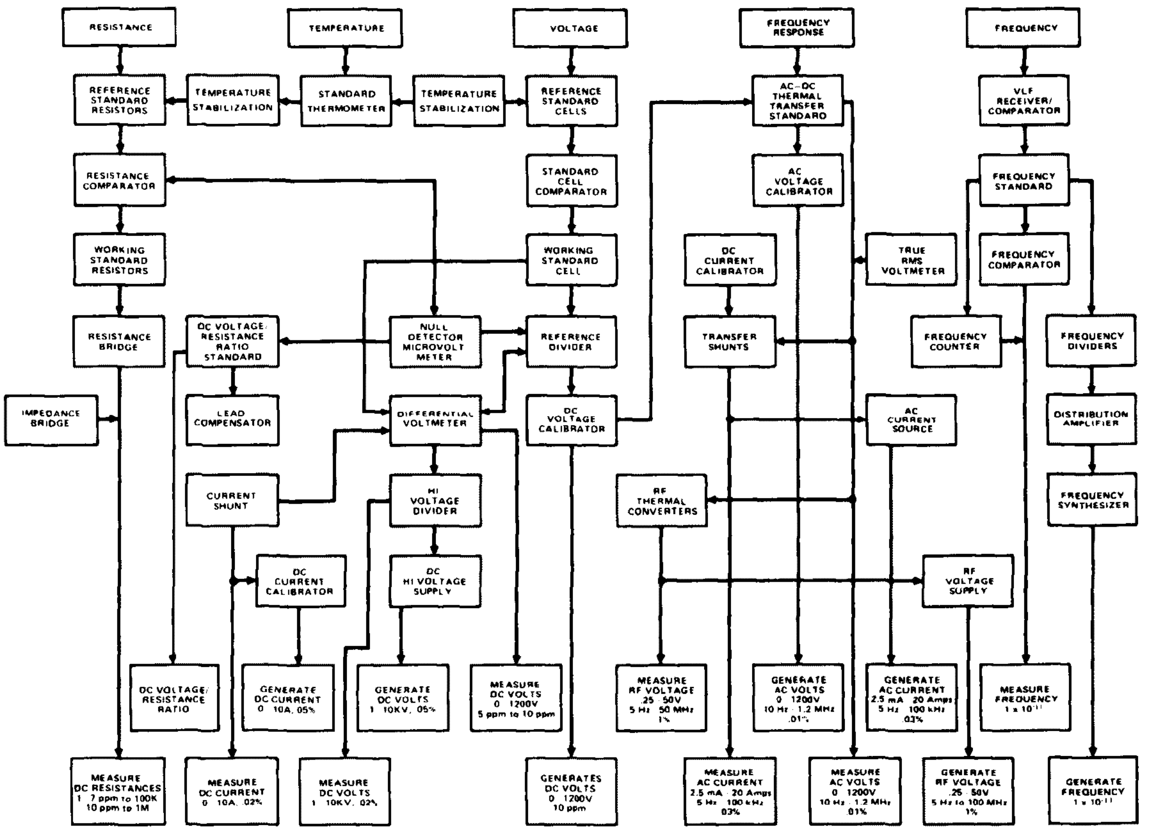

fig. 2. All measurements can be "traced" back to NBS National Measurement System.

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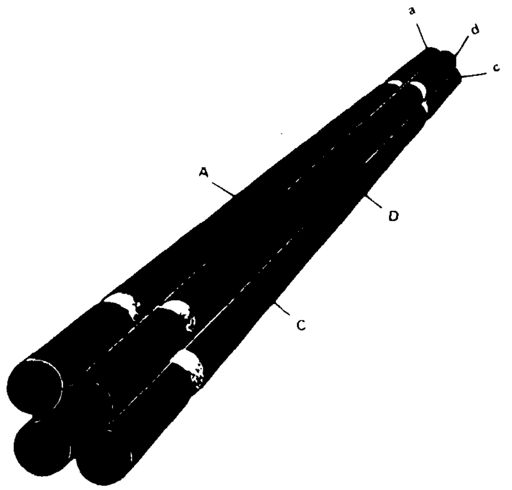

fig. 3. Basic structure of the first operating calculable capacitor, built at NBS in 1960. The calculated "cross capacitor" is the average of the capacitance (per meter of length) of the capacitor formed by rods A and C and that formed by B and D. When the "guard" sections a, are kept at the same voltage as A b, b at the same voltage as B, etc., then A, B, C, and D are very nearly as if parts of infinitely long rods.

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length is based on the meter, which is a set number of wavelengths of radiation of the Krypton 86 atom. Measurement of time is based on the second, which is defined as a certain number of cycles of radiation of the Cesium 133 atom in the NBS atomic clock.

In electronic communications, one of the most vital concerns is with the propagation of electromagnetic waves, the maximum velocity of which is limited only by the speed of light. Now that we know the exact value of length and time, it is possible — through the use of Maxwell's equations in electrodynamics and electrostatics — to precisely realize the relationships existing between permeability and permittivity of free space.* We can now derive one electrical standard from another (a technique mentioned earlier), and we do precisely this in the definition of the farad, the standard of electrical capacitance. Once we define the farad, all other electrical parameters fall into place.

defining the farad

The farad was really not very precisely defined until 1956 when two Australian physicists, Thompson and Lampard, contrived a technique of precisely measuring the lengths of two parallel metal rods in a special configuration, measuring the speed of light, and then calculating the value of the permittivity of free space. (See **figs. 3** and **4**.) Extensive work at the NBS followed, and these techniques were refined to the point where we now know the value of a farad to within ±2 ppm (parts per million). Now that the farad, the unit of capacitance, is known, determination of all other electrical units can follow. Impedance to the flow of AC is based on capacitive reactance formulas, and other electrical parameters such as current can be defined from this.

It is known that a quadrature or four-element bridge is an electronic circuit that can be used to compare the impedance of capacitors, inductors, and resistors. (Refer to the lower left portion of **fig. 5**.) A stable source of known ac frequency is all that is left to power the bridge. Since frequency is the inverse of time and since time is quite well defined by the atomic clock in the NBS facility in Boulder, Colorado, we are home free. Realizing also that there is one frequency, called the resonant frequency, at which inductive and capacitive reactances cancel one another, we can now define the henry from the farad, since the farad is already known. In addition,

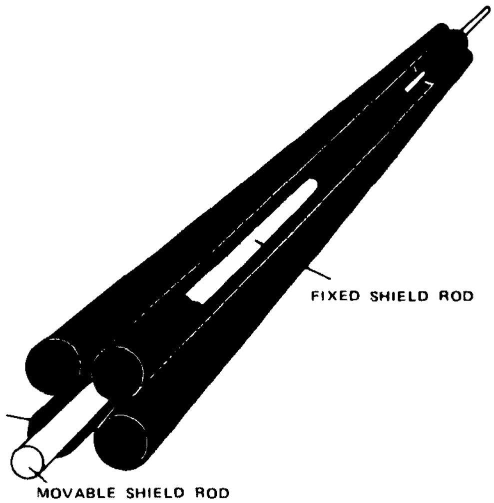

fig. 4. Basic structure of the variable cross capacitor used as standard of farad and ohm. The capacitance calculated is the change in cross capacitance when the movable shield rod is displaced through a measured distance along the axis. (Most of the rod D has been removed so the shield rod can be better seen. Note that the variable cross capacitor does not need guard sections. Omitted from this and the preceding figure is the metal shield that encloses the entire device.)

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*c *(velocity of light)* = $\frac{1}{\mu_0\epsilon_0}$ where $\mu_0\epsilon_0$ equals free space permeability and permittivity, respectively. **Editor**

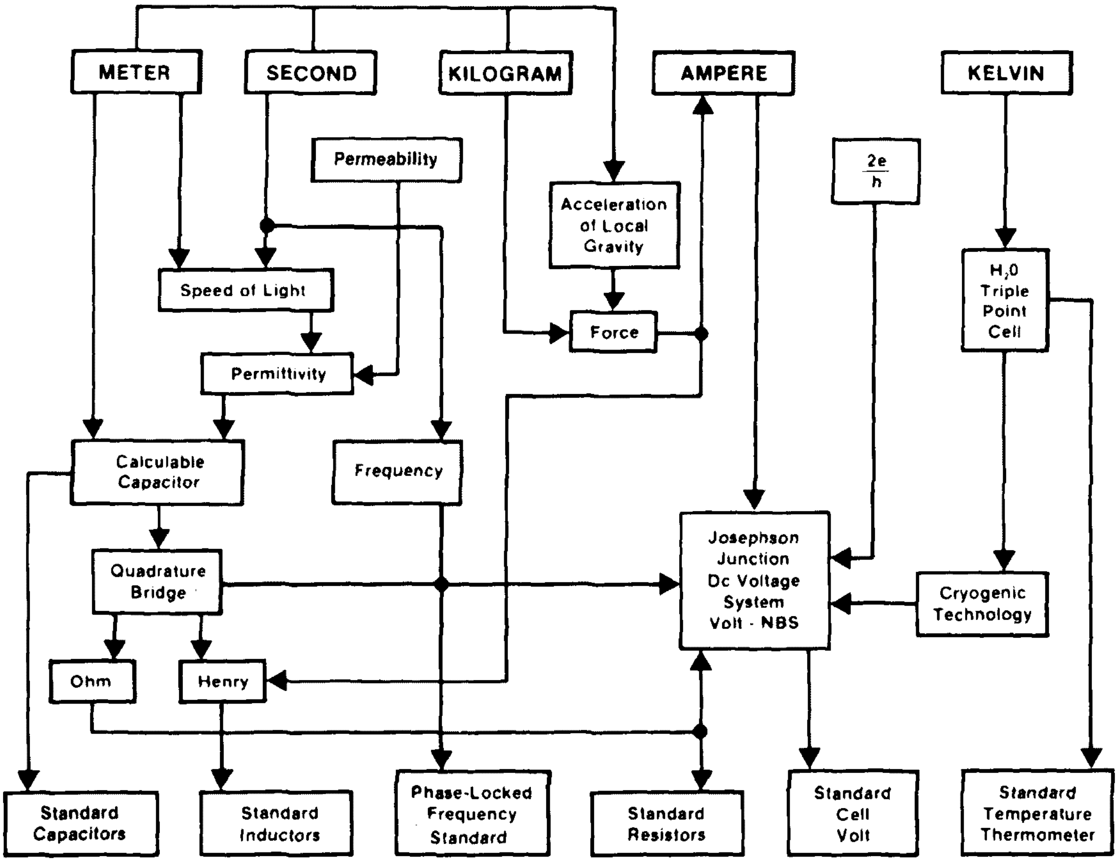

fig. 5. Derivation chart.

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through an AC-DC transfer technique, the ohm has been defined to better than ± 0.06 ppm. In other words, the ohm was *derived by calculating* the inductance of a carefully designed coil, taking into account all physical dimensions and the number of turns and type of wire used.

An ac source was then applied to the coil, and by knowing its frequency and being able to calculate the coil's inductance, the third quantity in the formula for inductance reactance (the resistance in ohms) was then derived.

the ampere

In Ampere's equation, current and its conductors are defined with respect to magnetic force. But first — in this hair-splitting business of defining absolute standards — *force itself* has to be very well defined. You will recall from physics that force is equivalent to mass times acceleration. The mass standard in the form of a kilogram bar of iridium and platinum already exists (in the BPIM, the French equivalent to our NBS). Next, we concern ourselves with acceleration — a special kind of acceleration, or the acceleration at local gravity at the site of the experiment. We derive acceleration by performing a drop test and measuring the distance traveled and the time elapsed.

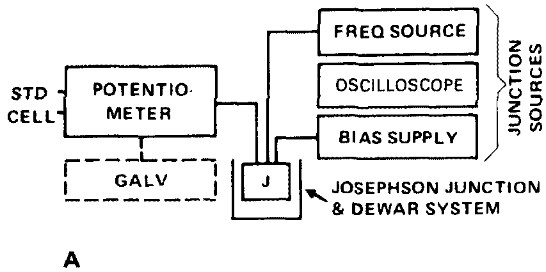
Having now established force and inductance from previous work on the quadrature bridge arrangement as a result of first knowing capacitance and time, we can proceed in defining the ampere absolutely by placing one free coil inside a fixed coil and attaching the free coil to the arm of a balance. Current is run through the coils; this counterbalances the gravational force, attracting a known mass which has been placed on the other arm of the balance. By opposing the downward force of local gravity on our known mass with magnetic fields resulting from the solenoid or magnetic "pulling" of our two coils and through Ampere's equation, an absolute determination of the unit of current, the ampere, is made.

the volt

The last unit to determine in the set of electrical standards is the volt. This unit can be determined in one of two ways. First, a very precise one-ohm resistor could have one ampere of current passed through it; as a result of Ohm's law, a one-volt drop would exist across this one-ohm resistor (see **fig. 6**). This very precisely determined volt would then be compared to some standard saturated cells and the results of this comparison would then be the nominal mean value assigned to this reference group. By having a number of standard saturated cells, the drifting of one cell to a few millionths of a volt above an actual volt would be offset by another cell drifting a few millionths of a volt in the other direction and collectively averaging all readings, a mean or average would then be obtained. But there are two problems with this technique, which the NBS used until most recently. First, the cells drifted; and second, since the volt was defined as a result of first determining the ampere, the problem was compounded by the practice of precisely redetermining the ampere only once or twice per decade.

The second method of determining the volt was much better and resulted from the work of a British physicist, Brian Josephson. In 1962, he published a paper entitled "Possible New Effects in Superconductive Tunneling" and won a Nobel Prize for what was later to be known as the "AC (alternating current) Josephson Effect." This discovery offers the computer industry semiconductor switching speeds approaching the speed of light and probably more importantly, offers a means by which the absolute volt can be very accurately determined.

To understand this wonderful discovery, let's first examine what a Josephson junction is and how it is used. In its simplest terms, the Josephson junction is a frequency-to-voltage converter. The frequency/voltage ratio is precisely equal to two times the elementary electron charge divided by Planck's Constant. The NBS places a resonant thin-film tunnel junction in an environment approaching absolute zero (0 degrees Kelvin or −273.15 degrees C) (see **fig. 7**). The result is a superconductor. Next, they interface the Josephson junction to an external voltage comparator which is connected to a Reference Standard Cell and then apply a stable dc supply voltage to the junction. This junction is irradiated with microwave energy from a Klystron tube that is phase-locked or precisely tuned by being locked to WWVB, the radio station at the NBS in Boulder, Colorado, that broadcasts the atomic time signals. (The atomic clock's quartz crystal oscillator is extremely accurate, with a short-term frequency stability of 1 part per 10^9.) The Josephson junction then generates a DC voltage between 1 and 10 millivolts, depending on the microwave frequency. The beauty of the system is that the resulting DC voltage has an accuracy approaching that of the Atomic Time Standard; furthermore, this can be reproduced anywhere in the world where WWVB or its equivalent can be received by radio and then processed for phase-lock control. The total system uncertainty in this voltage determining scheme is less than five parts per 10^8.



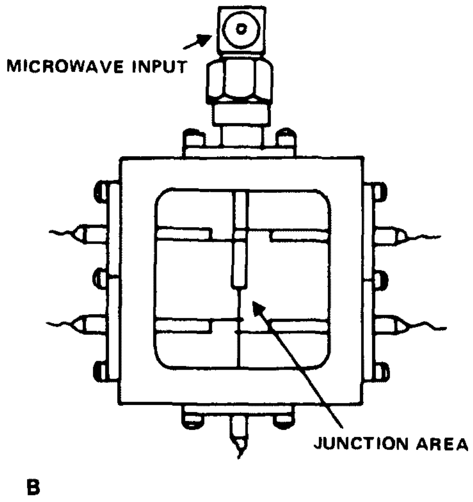


fig. 6(A) Block diagram of the overall voltage standard system, with a Josephson junction voltage reference for the potentiometer, which is used to calibrate standard cells. (B) Josephson junction device mounted in a microstripline enclosure (with the cover removed from the box).

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The resulting junction voltage is transferred up to the 1-volt level and then used as a check standard on a reference group of saturated cells. This small group then serves as a check on the working standards used for the Bureau's voltage calibration services.

ham radio

rf synthesizers for hf communications: part 3

Design and build a 5 to 5.5 MHz synthesizer

The first two parts of this series discussed the basic PLL. This third section applies the developed theory to the design and construction of a 5.000 to 5.500 MHz synthesizer that steps in 1 kHz increments, demonstrates synthesizer design, and provides some solutions to problems that may occur in synthesizer design.

synthesizer specifications

Loop performance specifications in a communications application such as an LO in a receiver or transmitter are phase noise, spurious content, and lock time. The best phase noise performance can be obtained with multi-PLL synthesizers[1] if space, complexity, and design time are not a problem. But because a simple single-loop synthesizer can be made more inexpensively and compactly than a multi-PLL synthesizer, design and construction of a simple single-loop synthesizer will be discussed here.

As a receiver LO, any spurs around the carrier can create spurious responses in the audio output. These spurs usually appear as reference feedthrough, power line (and harmonic) modulation of the VCO, and spurious oscillations in the VCO itself. Spurious oscillations in the VCO and power line modulation of the VCO are prevented by proper oscillator design, which includes good shielding and filtering. Reference feedthrough is largely a product of proper choice of loop parameters (f_n, ξ, etc.) and type of phase detector.

In the design loop, -70 dBc was chosen as a reasonable reference feedthrough limit, and -40 dBc for any spurs created by power line 60 Hz (and harmonics). Since spurs at 60 Hz (and odd harmonics) originate largely from the power lines coupling into the VCO, shielding is an effective technique for reducing them. 120 Hz spurs are usually the result of power supply ripple modulating the VCO and can be suppressed by filtering.

Lock time happens to be not only a function of loop bandwidth, but of frequency step size as well as the amount of frequency error, the designer considers acceptable for the "lock" state. Lock time for frequency synthesizers is usually specified as to within some Δf of the desired frequency before some specified time. Our loop must be within 1 Hz in less than 100 ms with a 1 kHz step in VCO frequency. A very rough equation for lock time for much less than 1 percent frequency error (relative to frequency step taken) is:

$$t_{LOCK} = 10/f_\beta$$

which is about 100 ms for a loop bandwidth of 100 Hz.

the phase detector

The design loop with an $f_{REF} = 1$ kHz requires a very narrow bandwidth to reduce reference feedthrough, while a wide loop bandwidth is required for a fast lock time. This design uses a relatively wide loop bandwidth ($f_\beta = 100$ Hz) for rapid lock time. To achieve the -70 dBc reference feedthrough requirement, a sample and hold (or an approximation of it) phase detector was used.

The phase detector is part of Motorola's MC145151 PLL IC. This phase/frequency detector goes tri-state (open circuit) with zero phase error at its inputs. As long as zero phase error is maintained, there will be no reference feedthrough. However, there is always a small amount of reference feedthrough, since zero phase error cannot be maintained indefinitely because of VCO drift, etc., which causes the phase detector to generate occasional corrective pulses (at an f_{REF} pulse rate).

programmable divider

The MC145151 PLL IC contains a programmable divider usable to at least 15 MHz at a supply voltage of 5 volts and a signal voltage of 500 mV peak-to-peak. The maximum division ratio is 16,383. It uses a straight binary input (not BCD), which facilitates linking to a computer, though not thumbwheel switches.

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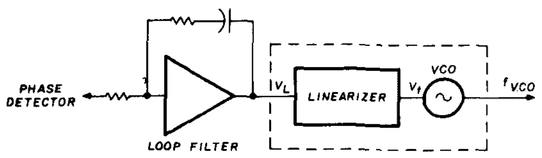

fig. 1. The linearizer that is used in conjunction with the VCO to achieve a linear f_{VCO} versus V_L transfer function.

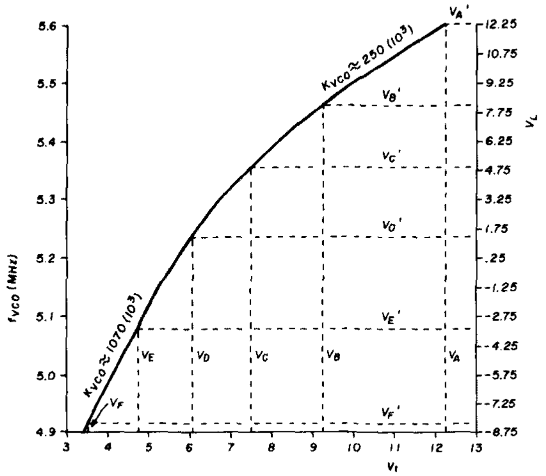

fig. 2. VCO transfer function showing f_{VCO} versus V_t and f_{VCO} versus V_L.

the VCO

In this narrow-loop bandwidth synthesizer design low spurious content and good phase noise performance directly relate to the VCO physical and electrical design. In addition it is also important that the VCO has a low frequency drift rate (to reduce reference feed-through), linear transfer function (to keep loop bandwidth and damping reasonably constant), and be well filtered and shielded (to reduce modulation of the VCO by power supply ripple and induced 60 Hz from power wiring).

VCO frequency drift can increase reference feedthrough by forcing the phase detector to generate correction pulses (which are not completely filtered by the loop filter) which in turn modulate the VCO. This is especially important when the reference frequency is close to the loop bandwidth, as is the case in this loop. Common contributors to frequency drift are the varactor diodes, toroidal core inductors in the resonator (if used), or excessive power in the resonator (unfortunately one of the requirements for a high signal-to-noise floor ratio for the oscillator).

Because a short lock time is required and f_{REF} is close to the loop bandwidth, then K_{VCO} must be fairly constant across the VCO frequency range in order to hold the loop bandwidth close to its designed value (and therefore reference feedthrough). Usually all that is required is to use the most linear section of the VCO tuning curve; its size dependent upon an acceptable value of nonlinearity. If the section of the VCO tuning curve used is small, this means a higher K_{VCO}, and the VCO will be more sensitive to unwanted signals on its tune line. The approach chosen here is to use a linearizer in combination with the VCO (**fig. 1**) to very closely approximate a linear VCO tuning curve (constant K_{VCO}).

the VCO linearizer

Fig. 2 shows the tuning curve for the VCO used in the design loop, with its 4.3:1 variation in K_{VCO}. Since this is significantly more than the 1.1:1 variation in K_N, (across the synthesizer frequency range) the effects of K_N on loop gain can be neglected in the following analysis.

With K_{VCO} the major cause of loop gain variations, its effects on f_n and ξ can be calculated:

$$\Delta_K = \sqrt{\frac{K_t\,(MAX)}{K_t\,(MIN)}} = \sqrt{4.3} = 2.07 \qquad (1)$$

Assuming that the loop bandwidth is 100 Hz and the damping factor equals 2 when the loop gain is greatest (i.e. K_{VCO} is at its maximum), then the effect of the 4.3:1 variation of K_{VCO} can be calculated:

$$f_n = \frac{f_\beta}{\sqrt{2\xi^2 + 1 + \sqrt{(2\xi^2 + 1) + 1}}}$$

$$= 100/4.25 = 23.53\text{ Hz} \qquad (2)$$

$$\text{And } f_n\,(MIN) = f_n\,(MAX)/\Delta_K$$

$$= 23.53/2.07 = 11.37\text{ Hz}$$

$$\xi(MIN) = \xi(MAX)\,/\,\Delta_K = 2/2.07 = 0.97$$

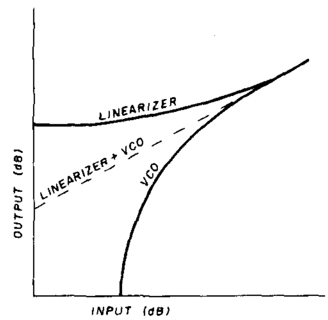

fig. 3. Adding the linearizer transfer function to that of the VCO to achieve a linear transfer function.

$$f_\beta = f_n \sqrt{2\xi^2 + 1 + \sqrt{(2\xi^2 + 1)^2 + 1}}$$

$$= 11.37\text{ Hz}\,(2.44) = 27.74\text{ Hz} \qquad (3)$$

which results in a 100/27.74 (3.6:1) change in loop bandwidth with its resulting effect on lock time.

The linearizer is designed to have a response that is the inverse of the VCO (nonlinear) transfer function so that when the linearizer is added to the VCO it will have the effect of producing a linear VCO tuning curve with constant K_{VCO} (**fig. 3**).

The linearizer is implemented (**fig. 4**) using resistor/diode sections[2] and a voltage divider network in order to individually modify the slope of the linearizer transfer function as each resistor/diode section is forward biased. (As V_L decreases, successive diodes are turned on modifying the slope of the V_L versus V_t curve).

The method used to design the linearizer required the following steps:

1. plotting K_{VCO} versus V_t and then determining the VCO range to be linearized,

2. calculating the quantity of resistor/diode sections required for the maximum allowed variation in K_{VCO} (realizing that this is only a segmented approximation of the required transfer function to achieve a linear VCO tuning curve),

3. calculating the voltages for the voltage divider,

4. calculating the resistor values for the resistor/diode sections.

The VCO frequency range to be linearized is roughly 4.92-5.6 MHz which allows for some drift in frequency. The K_{VCO} at these two frequencies are 1070 (10^3) and 250 (10^3) rad/sec/volt. The number of resistor/diode sections needed for the linearizer are approximately:

$$n \approx \frac{\log \dfrac{K_{VCO}(MAX)}{K_{VCO}(MIN)}}{\log \Delta K_{VCO}} \qquad (4)$$

where ΔK_{VCO} is the allowed variation in K_{VCO} after it is linearized. For the design loop, it was decided that K_{VCO} will be allowed to change by only 30 percent so

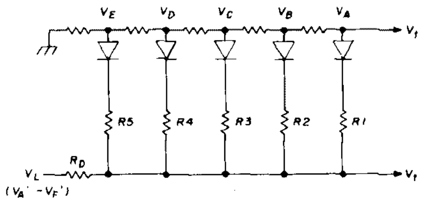

fig. 4. Linearizer using multiple diode/resistor sections to approximate the desired transfer function.

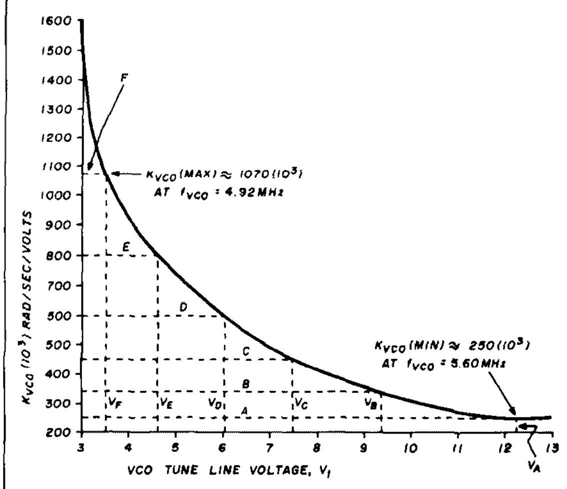

fig. 5. Plot of K_{VCO} versus V_t.

ΔK_{VCO} = 1.3. (Once again, neglecting K_N, if K_{VCO} is the only source of loop gain variation, it varies only by 30 percent, then f_n and ξ will vary only by the square root of this, or 14 percent. This is acceptable.)

Therefore the number of resistor diode pairs needed are:

$$n = \frac{\log \dfrac{1070\,(10^3)}{250\,(10^3)}}{\log 1.3} = 5.54$$

It is decided to use 5 resistor/diode pairs instead of 6 even though this means a slightly higher error. This error is approximately:

$$\Delta K_{VCO} = 10^{\dfrac{\log \frac{K_{VCO}(MAX)}{K_{VCO}(MIN)}}{n}}$$

$$= 10^{\frac{\log 4.3}{5}} = 1.337 \qquad (5)$$

Knowing ΔK_{VCO} the voltage divider voltages can be calculated. In **fig.** 5 a mark for every 33.7 perent increase in K_{VCO} (starting at the low end of the K_{VCO} plot, or 250 (10^3) in the example) corresponds to a voltage on the VCO tune line, V_t. If all the calculations are computed correctly, then the last computed point should correspond to the maximum K_{VCO} of the range over which the linearizer was designed to work ($1070 \cdot 10^3$).

$$K_{VCO}(B) = 1.337\,K_{VCO}(A)$$
$$= 1.337 \cdot (250 \cdot 10^3) = 334.25 \cdot 10^3$$

$$K_{VCO}(C) = 1.337\,K_{VCO}(B)$$
$$= 1.337 \cdot (334.25 \cdot 10^3) = 446.89 \cdot 10^3$$

$$K_{VCO}(D) = 1.337\,K_{VCO}(C)$$
$$= 1.337 \cdot (446.89 \cdot 10^3) = 597.5 \cdot 10^3$$

$K_{VCO}(E) = 1.337\, K_{VCO}(D)$

$= 1.337 \cdot (597.5 \cdot 10^3) = 798.8 \cdot 10^3$

and for a check

$K_{VCO}(F) = 1.337\, K_{VCO}(E)$

$= 1.337 \cdot (798.8 \cdot 10^3) = 1068 \cdot 10^3$

which is very close to $1070 \cdot 10^3 = K_{VCO}(MAX)$.

The corresponding trip voltages (where the resistor/diode combination starts conducting) are:

$V_A = 12.25$ volts	$V_D = 6.05$ volts
$V_B = 9.35$ volts	$V_E = 4.60$ volts
$V_C = 7.70$ volts	$V_F = 3.50$ volts

The voltage divider component values are chosen such that the current through them is much higher than the maximum current through the diodes to keep the voltages V_A through V_E constant.

To calculate the resistor values R1 through R5 and R_D, the voltage V_L that corresponds to the voltage V_t (especially at the trip voltages V_A through V_E) must be determined. This is done by letting V_A' equal V_A and then linearly scaling the voltages V_B' through V_F' such that these voltages are always less than their corresponding voltages V_B through V_F (**fig. 2**). This means that V_L will swing through a larger range of voltages (-9.25 to 12.25) than V_t, (3.5-12.25) as detailed below:

$V_A' = 12.25$	$(V_A' = V_A)$
$V_B' = 8.2$	$(V_B' < V_B)$
$V_C' = 4.9$	$(V_C' < V_C)$
$V_D' = 1.4$	$(V_D' < V_D)$
$V_E' = -3.25$	$(V_F' < V_E)$
$V_F' = -9.2$	$(V_F' < V_F)$

By assuming perfect diodes (i.e., no forward voltage drop), resistors R1 through R5 can be determined. To keep the current through the diodes at a minimum R_D should be fairly high (46.4k chosen here). By knowing what voltage the linearizer requires at its input (V_L) to generate a corresponding voltage at the output, (V_t) (**fig. 2**) resistors R1 through R5 can be calculated (**fig. 6**).

Power supply ripple is kept off the VCO tune line by using a Darlington transistor with a large RC time constant in its base that buffers the voltage to the linearizer voltage divider.

Ideally, the linearizer provides a perfectly constant K_{VCO} across the VCO tuning range. Because the linearizer is synthesized using linear segments (and these linear segments are really nonlinear due to the imperfect diodes), there will still be a non-constant K_{VCO} (non-linear VCO tuning curve), though vastly improved. **Fig. 7** shows the measured K_{VCO} versus V_L using the linearizer and is compared to **fig. 5** to show the improvement. The error appears to be very close to the calculated value of 33.7 percent.

loop filter

With K_{VCO} almost constant and K_ϕ and K_N defined, the loop filter components can now be determined. Looking at **fig. 7**, the minimum and maximum values for K_{VCO} are:

$$K_{VCO}(MAX) = 250 \cdot 10^3$$
$$K_{VCO}(MIN) = 190 \cdot 10^3$$

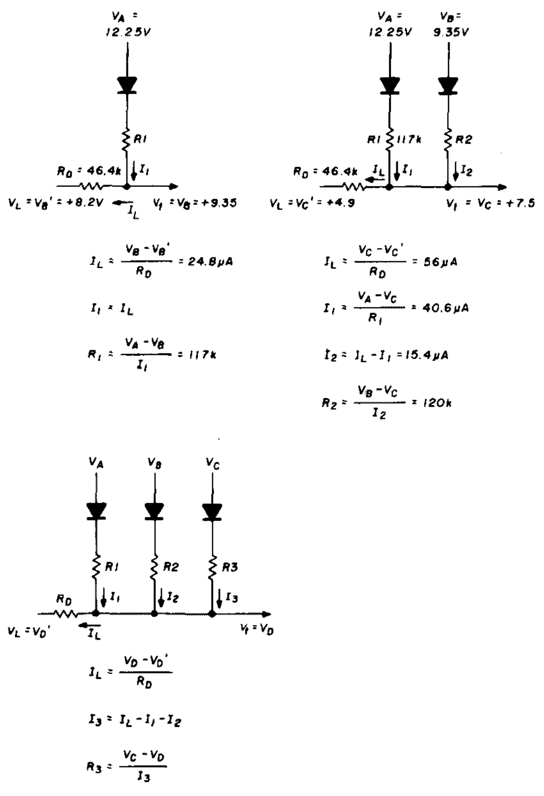

fig. 6. Calculating the slope setting resistor values for the linearizer.

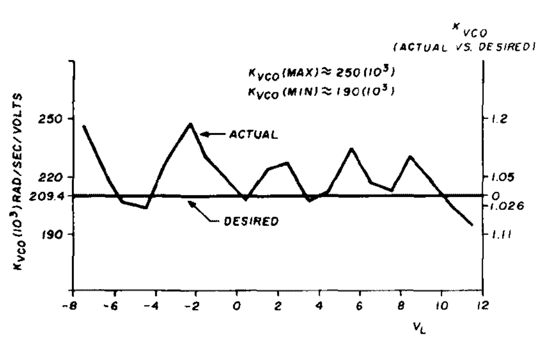

fig. 7. Plot of K_{VCO} versus V_L using linearizer.

The values for $K_N(MAX)$ and $K_N(MIN)$ were as stated before:

$K_N(MAX) = 200 \cdot 10^{-6}$
$K_N(MIN) = 181.82 \cdot 10^{-6}$
$K_\phi = 0.398$ volts/radian

Assuming $K_{VCO}(MAX)$ occurs at $K_N(MAX)$ and $K_{VCO}(MIN)$ at $K_N(MIN)$ (for worse case open loop gain variation) then:

$K_t(MAX) = 19.9$
$K_t(MIN) = 13.75$

Variations in f_n and ξ are:

$$\Delta_K = \sqrt{\frac{19.9}{13.75}} = 1.2$$

It was decided to have a damping factor of approximately one, but since it will vary by $\Delta_K = 1.2$ the lower value of ξ will be set at $\xi = 0.95$ and the maximum value will then be $\xi = 0.95\,(1.2) = 1.14$, which, as discussed before, corresponds to the maximum loop bandwidth of 100 Hz. Using this information both $f_n(MAX)$ and $f_n(MIN)$ can be calculated:

$$f_n(MAX) = \frac{f_\beta(MAX)}{\sqrt{2\xi^2 + 1 + \sqrt{(2\xi^2 + 1)^2 + 1}}}$$
$$= 100/2.708 = 36.92 \text{ Hz}$$

and

$f_n(MIN) = f_n(MAX)/\Delta_K = 36.92/1.2 = 30.77$ Hz

With maximum values of damping, loop natural frequency, and loop gain known, the two time constants for the loop filter can now be determined.

$$t_1 = \frac{K_t}{(\omega_n)^2}$$

$$t_2 = \frac{2\xi}{\omega_n}$$

At $f_{VCO} = 5.00$ MHz the loop gain is at its maximum corresponding to maximum values for f_n, f_β, and ξ. Using these maximum values the time constants are:

$$t_1 = \frac{19.9}{(2\pi 36.92)^2} = 369.8\ \mu s$$

$$t_2 = \frac{2(1.14)}{2\pi 36.92} = 9.828\ ms$$

Using minimum values of K_t, f_n, and ξ should produce the similar results:

$$t_1 = \frac{13.75}{(2\pi 30.77)^2} = 367.8\ \mu s$$

$$t_2 = \frac{2(0.95)}{2\pi 30.77} = 9.83\ ms$$

Although the loop filter components could now be calculated using t_1 and t_2, there are two areas that will restrict the values of R1 and C1 and should be examined (**fig. 8**). These areas are leakage currents and phase detector output current.

With the loop in lock the phase detector is in an open-state mode and does not "source" or "sink" current to the integrator, thereby allowing the integrator to hold its charge. Therefore the voltage out of the loop filter is constant. However, leakage currents will either add or subtract charge from the integrator capacitor causing the loop filter output voltage to the VCO to vary, changing its frequency. The loop then corrects for this frequency shift during the next reference frequency clock cycle. This means that a signal, at a frequency equal to the reference frequency and of an amplitude dependent on the leakage current, will be on the VCO tune line and thus modulate it (reference feedthrough).

Given the equation for the integrator: $\Delta V_{OUT} = I/C\,(\Delta t)$, where I is the leakage current, $\Delta t = 1/f$, and C is the integrator capacitor ($C1$), C is really the only variable available to the designer. (f_{REF} is fixed, and I is the leakage current of both the phase detector in its open circuit mode and the bias currents of the (LF-356) op-amp used in the integrator. The leakage current for the phase detector is 1 nanoampere and for the op-amp is about 10 pico ampere. The phase detector leakage predominates and so the op-amp leakage can be neglected.)

With maximum acceptable spur level of −70 dBc, and knowledge of $K_{VCO}(MAX)$, the maximum reference feedthrough voltage on the VCO tune line can be approximated, then the smallest value of C1 for that reference feedthrough voltage can be calculated (given the leakage current).

$$V_L\,(p\text{-}p) = \frac{8\pi f_m 10^{\frac{P_{SSB}}{20}}}{K_{VCO}(MAX)}$$

$$= \frac{8\pi\,(1kHz)\,10^{\frac{-70}{20}}}{250 \cdot 10^3}$$

$$= 31.8\,\mu V(p\text{-}p) \approx 32\,\mu V(p\text{-}p)$$

where V_L = (p-p) reference feedthrough signal voltage
P_{SSB} = single sideband spur level in dBc
f_m = reference feedthrough signal frequency

The minimum value of $C1$ is:

$$C1(MIN) = (I)\,\frac{\Delta t}{\Delta V} = (I)\,\frac{1/f_{REF}}{\Delta V_L}$$

$$= 1\text{ nA}\left(\frac{1\text{ ms}}{32\ \mu V}\right) = 0.031\ \mu F$$

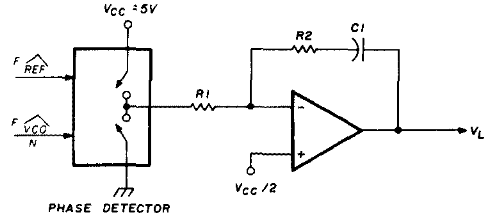

fig. 8. Phase detector (in its open-circuit mode) plus loop filter.

where I is the leakage current and V is the calculated reference feedthrough voltage for a −70 dBc spur level.

The smallest value of R1 is determined by the maximum current the phase detector can develop which, for the MC145151 is about 500 μA. With the maximum phase detector voltage of approximately +5V and the other extreme at (a virtual) 2.5V then:

$$R1(MIN) = (5-2.5)/500\mu A = 5000 \text{ ohms}$$

The loop filter components can now be calculated by letting R1 = 5000 and solving for C1:

$$C1 = t_1/R1 = 367.8\,\mu s/5000 = 0.0736\,\mu F$$

which fortunately happens to be greater than the minimum value for C1 calculated before. Letting C1 take on a more practical value, *C1 = 0.068 μF,* R1 can be recalculated:

$$R1 = 367.8\,\mu s/0.068\,\mu F = 5410 \text{ ohms}$$

Calculating for R2:

$$R2 = t_2/C1 = 9.83\,ms/0.068\,\mu F = 144.5K$$

or 147K, a common value in 1 percent resistors. Both R1 and C1 are within the desired limits.

For prefiltering ahead of the loop filter R1 is split into two series resistors and a capacitor is tied between their common points and ground.

A unity gain inverter using a low noise op-amp follows the loop filter for required signal inversion

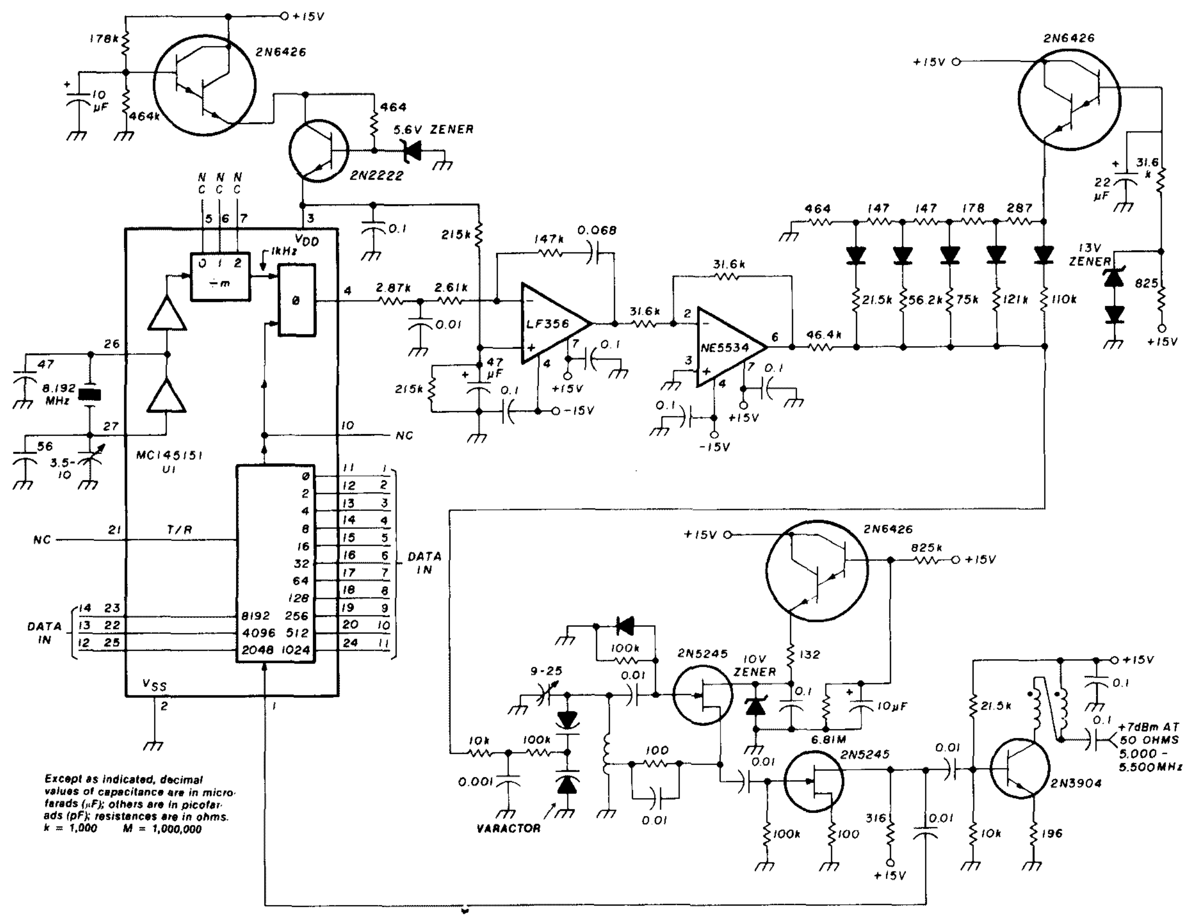

fig. 9. Synthesizer schematic. The physical dimension of the completed synthesizer is 2 by 4 inches.

for negative feedback in the phase-locked loop. Another filter pole is added to the loop by the addition of a 0.01 μF capacitor across the inverters' feedback resistor. Like the pre-filter ahead of the loop filter, this pole is located at a frequency well outside the loop bandwidth.

loop performance

This synthesizer (**fig. 9**) performed according to specifications immediately after completion. The only problem was excessive 60 Hz (and its odd harmonics) spurs which were solved with proper shielding. A Hewlett-Packard HP8568 spectrum analyzer was used to measure phase noise and spurious responses. **Fig. 10** uses a wide-frequency sweep to show the general "cleanliness" of the synthesizer output for large frequency offsets from the carrier. The noise pedestal around the carrier is that generated by the spectrum analyzer.

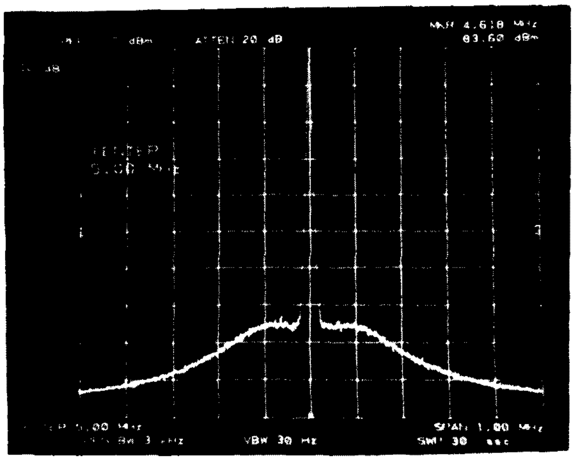

fig. 10. Phase and amplitude noise spectrum of the synthesizer for large offset frequencies ($\pm$500 kHz).

Fig. 11 is a SSB phase (and amplitude) noise plot of the synthesizer output using a 5-kHz linear sweep. The spurs are caused by the phase detector; examined closely, they can be seen to be two noise peaks centered around where the reference feed-through spur should be. Upon close examination the reference feed-through was greater than -70 dBc, as designed. Unfortunately temperature effects on phase detector leakage current were not taken into account (as well as the variations in leakage current among MC145151's) and the -70 dBc spur requirement is exceeded as the MC145151 heats.

Fig. 12 is a narrow spectrum view of the synthesizer using a 500 Hz sweep. This shows what reasonable shielding and filtering can do to minimize spurs. The 60 Hz spurs, originally only -30 dBc,

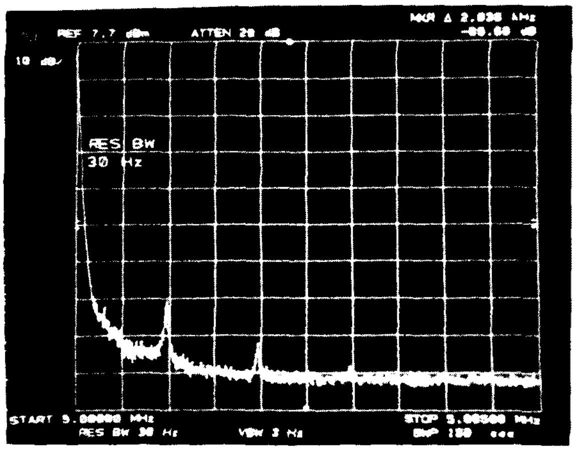

fig. 11. SSB phase and amplitude noise spectrum for up to 5 kHz offset. Spurs are actually two noise peaks centered around the reference feedthrough spur.

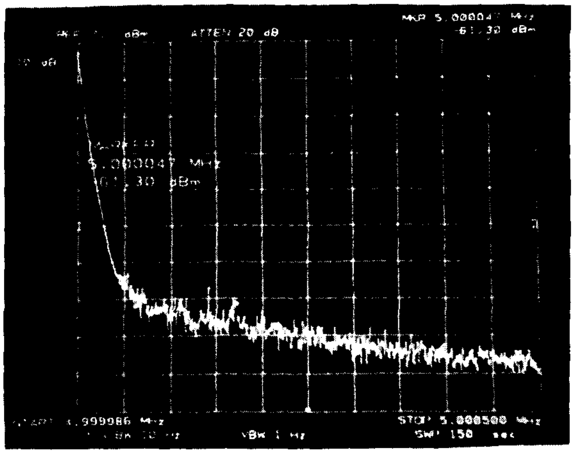

fig. 12. Close in phase and amplitude noise spectrum of synthesizer showing no perceptible 60 Hz or harmonic related spurs.

were reduced an additional 15 dB by optically shielding the varactor diodes, and by more than 30 dB through magnetic shielding of the synthesizer. The 120 Hz spurs were taken care of by using a Darlington transistor as an active filter for supply lines to sensitive parts of the synthesizer (i.e., VCO, linearizer, etc.).

Fig. 13 is a phase noise plot of the synthesizer at 5 MHz. Since the phase noise past the loop bandwidth (offset frequencies greater than f_β = 100 Hz) is equivalent to that of the open loop VCO, while loop bandwidth and damping are held fairly constant, then the phase noise plot taken at any other frequency within the synthesizer range will be essentially identical.

The lock time was examined by changing the programmable divider by one (Δf_{VCO} = 1 kHz) and measuring the voltage settling time at V_L for a

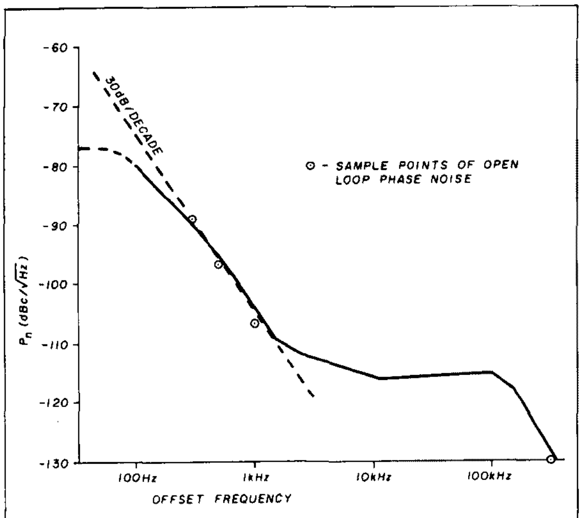

fig. 13. Phase plus amplitude noise for the synthesizer at 5.000 MHz.

roughly 1 percent frequency error. This turned out to be about 30 ms, much less than the $10/f_\beta$ calculated (which shows just how rough an estimate of $10/f_\beta$ is).

using the Bode plot

Another method used in designing the loop filter, as well as a check and debugging aid, is by using the Bode plot. This is done by plotting the transfer function of $K_t = K_\phi\ K_{VCO}\ K_N$, then plotting the desired transfer function of $K_\phi\ K_F\ K_{VCO}\ K_N$. Subtracting the two gives the required transfer function of the loop filter from which the time constants for it can be derived.

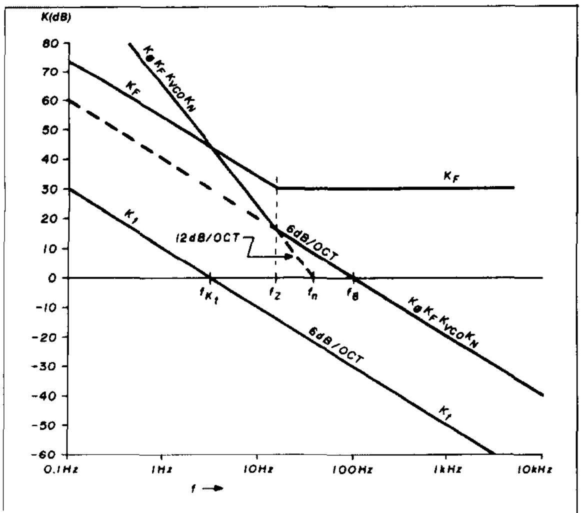

fig. 14. Bode plot of synthesizer open-loop gains to help determine loop filter time constants.

In the example loop, $K_t = 19.9$ (at $f_{VCO} = 5.0$ MHz). A plot of K_t would have a 6 dB/octave slope crossing zero at a frequency of:

$$f = K_t/2\pi = 19.9/2\pi = 3.2\ \text{Hz}$$

The desired open-loop response, $K_\phi\ K_F\ K_{VCO}\ K_N$ (**fig. 14**) crosses the desired bandwidth $f_\beta = 100$ Hz with a 6 dB/octave slope at unity gain, then intersects and assumes the 12 dB/octave slope of a line that crosses unity gain at:

$$f_n = \frac{f_\beta}{\sqrt{2\xi^2 + 1 + \sqrt{(2\xi^2 + 1)^2 + 1}}} = 39.4\ Hz$$

where the desired value of damping is one.

Subtracting K_t from $K_\phi\ K_F\ K_{VCO}\ K_N$ (the desired total open-loop response) gives K_F as shown in **fig. 14**. The 6 dB/octave slope of K_F levels off at f_Z at A = 30 dB. The resistor ratios for the loop filter (**fig. 8**) are:

$$R2/R1 = Y = 10^{A/20} = 10^{+30/20} = 31.62$$

The time constants are:

$$t_2 = 1/(2\pi f_Z) = 1/(2\pi\ 15.5\ \text{Hz}) = 10.27\ \text{ms}$$

$$t_1 = t_2/Y = 10.27\ \text{ms}/31.67 = 324.2\ \mu\text{s}$$

As expected, these are very close to the earlier designed values.

This plot could further include all poles and zeros in the loop to spot potential problems (i.e., too much phase shift where it may cause loop instability).

conclusion

This synthesizer should work fairly well as an LO for a receiver or transmitter if a 1 kHz step is small enough. For a synthesizer requiring smaller frequency steps a different configuration may be used, in-loop mixing with a VXO being a good choice.

It is important to realize that this synthesizer wanted a fast lock time with low reference feedthrough, and paid a rather high price — i.e., inclusion of the linearizer — for it. If a longer lock time is acceptable (lower f_β to accommodate variations in loop gain), the linearizer can be left out of the circuit, simplifying the design. However, in the application for which this synthesizer will be used (frequency scanning), a rapid lock time is desirable. It also must be mentioned that as f_β decreases, the loop will have increasing difficulty in tracking mechanical vibrations, and even its own VCO drift.

references

1. Craig Corsetto, WA6OAA, "RF Synthesizers for HF Communications: part 2," *ham radio*, September, 1983, page 48.
2. William Egan, *Frequency Synthesis by Phase Lock*, pages 55-58, John Wiley and Sons, 1981.

ham radio

vertical phased arrays: part 4

Feed network design using L-match circuits, π and tee coax-equivalent circuits

Previous articles of this series on vertical phased arrays[1,2,3] concentrated on the design of the physical aspects of arrays: element length, radius, spacing and ground planes. The latest article[3] dealt with electrical measurements of the arrays and calculation of driving-point impedances. Knowing the required drive current amplitude and phase for each element of the array pattern selected, and knowing the measured values of self- and mutual impedances, we can calculate the driving-point impedance of these elements. The importance of this cannot be over-emphasized; because of mutual impedance effects between elements, driving-point impedances of elements in an array are not fixed entities. Each element's driving-point impedance depends upon the amplitude and phase of the drive currents — not just upon its own drive current, but upon the amplitude and phase of the drive current of every element in this array. A change of the current amplitude or phase to any element results in a driving-point impedance change of *every* element together with a change in *all* current amplitudes and phases. Examination of the set of simultaneous equations defining these driving-point impedances illustrates this relationship:

$$Z_1 = E_1/I_1 = Z_{11} + I_2Z_{12}/I_1 + I_3Z_{13}/I_1 + \cdots + I_n/Z_{1n}/I_1 \qquad (1)$$

$$Z_2 = E_2/I_2 = I_1Z_{12}/I_2 + Z_{22} + I_3Z_{23}/I_2 + \cdots + I_nZ_{2n}/I_2$$

$$Z_n = E_n/I_n = I_1Z_{1n}/I_n + I_2Z_{2n}/I_n + I_3Z_{3n}/I_n + \cdots + Z_{nn}$$

where $Z_1, Z_2, \cdots, Z_n$ are element driving-point impedances

$E_1, E_2, \cdots, E_n$ are element impressed voltages

$I_1, I_2, \cdots, I_n$ are element drive currents

$Z_{11}, Z_{22}, \cdots, Z_{nn}$ are element self impedances

$Z_{12}, Z_{13}, Z_{23}, \cdots, Z_{1n}, Z_{2n}, Z_{3n}$ are mutual impedances between pairs of elements

(All terms can be complex.)

For example, suppose the drive current I_2 to element 2 changes. Since I_2 appears in the equation for every element, all driving-point impedances, currents and voltages are affected. The self and mutual impedances do *not* change, even though it is the mutual impedances that cause this interaction.

An array is a coupled system, automatically adjusting to any change with a new set of currents and phases, which again simultaneously satisfies all the equations. There is an infinite number of such solutions, but *only a few* result in useful array patterns. A feed network must be designed that when connected to the terminals of each element, applies the proper voltage amplitude and phase, causing the required drive currents to flow. If this condition is met, then all conditions are met. (It may now be clearer why I have been emphasizing the importance of physical and electrical symmetry of the elements.) As the array direction is switched, each port of the feed network continues to "see" the same driving-point impedance it was designed for, even though each port is now feeding a different element. Exact symmetry is probably the most difficult condition to meet in practice because it depends upon more than just simple duplication of physical elements; it also depends on duplication of the environment adjacent to each element: for example, ground planes or other nearby conductors that might act as antennas.

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feed networks

Just as there are an infinite number of solutions to the set of equations defining the driving-point impedances of an array, there are almost as many ways to design networks fitting the one solution required. Some designs are better than others, resulting in more bandwidth for usable F/B (front-to-back) ratio or low SWR. As a general rule, the simpler (that is, the fewer stages in the network), the better, but there are exceptions.

For superior F/B performance, for all network designs, the designer must know the driving-point impedance of each element. In this respect, vertical phased arrays are more critically affected by element variations than Yagis are by height variations. For multiple-element arrays, cookbook recipe duplication attempts are almost guaranteed to miss optimum current drive conditions by 10 percent. *This is enough to reduce the F/B performance by 50 percent* or more.

basic design objectives

Some basic design decisions must first be made: what type of circuit elements should be used? Should the design have the objective of a 1:1 SWR match to the array feedline? Feed networks may be devised using coaxial cables as circuit elements. Although simple in construction, there may be technical and cost drawbacks. Series or stub coaxial cable sections, provided one has a wide enough selection of different characteristic impedances and lengths, could be used for a network that matches to the feedline. The cost would be high and bandwidth narrower than the array's intrinsic F/B capability.

As a special case, for arrays operating with 90-degree current phase multiples and equal current amplitude ratio element pairs,[3] an approach suggested by W7EL[4] makes use of the unique characteristic of a 1/4-wavelength line to produce a constant 90-degree phase displacement between input voltage and output current, independent of the load termination. If two such lines are connected to a common feedpoint, equal current will flow into the loads regardless of their termination impedances. The current phase displacement between the two loads is 0 degrees, but this may be changed to 180 degrees by insertion of a 1/2-wavelength line in series with one of the lines. (The phase displacement of a 1/2-wavelength line also is independent of the load impedance.) A 90-degree current phase displacement, as was pointed out earlier in this series,[1] cannot be obtained by insertion of a 1/4-wavelength line when the termination is reactive. Therefore, instead of inserting an additional 1/4-wavelength line, a lumped-constant phase correction circuit based upon the calculated driving-point impedances is inserted. It provides a drive current phase of 90 degrees and the correct amplitude at the element(s). The input voltage amplitude and phase to the correction circuit must be designed to be the same as that of the common connection point of the array. SWR of the array can be minimized by proper choice of the characteristic impedance of the coaxial cable feeder lines but cannot be designed for a 1:1 condition. This approach, based upon this unique characteristic of 1/4-wavelength lines, which are also the element feedlines, is limited to arrays where this length is able to physically reach the directional switch. If not possible, a further 1/2-wavelength line has to be added to each feedline to maintain the basis of the design concept.

This article provides sufficient information to enable the reader to design a feed network for any conceivable array. There are no restrictions on array spacings, current amplitude ratios or phase displacements. The elements must be alike, but they need not be resonant. Conventional or not, if the array you've been able to fit onto your property has a useful pattern, a feed network can be designed to drive it at the needed conditions. Such versatility requires complete design freedom — freedom to use any characteristic impedance, to transform to any input resistance, regardless of reactive load impedances. Coax is an excellent means for transmission of RF energy between physically separated points. If it also fills a role as a specific circuit element, so much the better. But as a circuit element where the physical spacing does not require it, coax is confining; with only two characteristic impedance choices commonly available (there are perhaps two more, but neither is easy to find), one is constantly making compromises and designing around this limitation. Furthermore, ease of circuit adjustment is not notable. On the other hand, π and tee coax-equivalent lumped-constant circuits may be designed for any exact characteristic impedance or any electrical length, whether lagging or leading phase, and are easily adjusted. And surprisingly, low impedance lumped constant circuits of the same levels as coaxial transmission lines display comparable characteristics, even when designed for fairly large single increments of phase displacement. **Table 1** compares coaxial cable with a 45-degree π circuit cascaded with a 45-degree tee circuit operated as a 1/4-wavelength transformer. Off-frequency phase variation and development of input reactance compares very favorably with coaxial transmission line.

Rounding out the list of network building blocks are the shunt and series input L-match transforming circuits. Included are the two special cases of this circuit, where the series or shunt branch is absent, which I will call a Parallel and a Series impedance circuit, respectively. **Figs. 1A** through **1F** are schematics of all of these circuits.

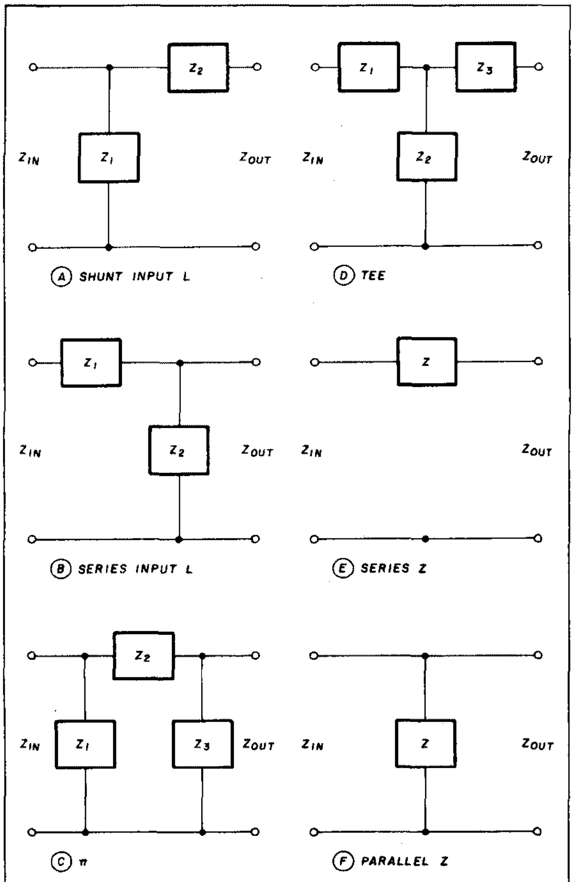

fig. 1. Basic network "building blocks."

why a 1:1 SWR?

For multi-element arrays the objective of a 1:1 match to the array feedline does not stem from an obsession with SWR. *A low SWR provides no significant measure of an array's efficiency or usable F/B bandwidth.* Designing for an SWR of 1:1 simplifies network design calculations and electrical tests. However, the real value is the instant array condition conveyed every time an SWR measurement is made. A failing relay in a directional switch or a network malfunction is quickly detected, even if this circuit is to an element requiring very little power. Such a failure may raise the SWR from 1:1 to 1.1:1, for example, while the same failure in an array normally showing 2:1 will not be noticed. At "smoke" test time it removes uncertainty; a 1:1 SWR represents an unambiguous confirmation of the accuracy of array measurements, network design, and construction.

As an illustration of the step-by-step design procedure for an array feed network, I will use the popular 2-element array. The same array will be used to show the error that arises in many 2-element vertical array feed arrangements. From Part 3 of this series,[3] the driving-point impedances of two 1/4-wavelength resonant elements spaced 1/4-wavelength apart with unity current ratio and 90 degree phase displacement, are:

Element 1	**Element 2**
$Z_1 = 21.4 - j15$	$Z_2 = 51.4 + j15$
$I_1 = 1 \angle 0^\circ$	$I_2 = 1 \angle -90^\circ$

Assuming 50 ohm 1/4-wavelength feedlines, using a Smith Chart or by calculation, the element driving-point impedances rotated to the input ends of the feeders are:

$$Z_1 = 78.3 + j54.91 \qquad Z_2 = 44.82 - j13.08$$

At element drive conditions of 1 ampere with a phase displacement of −90 degrees between elements, the voltages and currents that must be applied to the inputs of these feeders in polar form are:

$$I_1 = 0.52 \angle 55.0^\circ \qquad I_2 = 1.07 \angle 16.3^\circ$$
$$E_1 = 50 \angle 90^\circ \qquad E_2 = 50 \angle 0^\circ$$

Notice that the current phase change in the two equal 1/4-wavelength feeders are 55 degrees and 106.3 degrees (90 degrees + 16.3 degrees). Next a 1/4-wavelength 50-ohm delay line is added to the feedline from element 2. Rotating the impedance to the end of the delay line we find these conditions:

$$Z_2 = 51.4 + j15$$
$$I_2 = 1 \angle 90^\circ$$
$$E_2 = 53.54 \angle 106.3^\circ$$

These are the conditions that must exist at the input ends of the feeders from each element for the assumed drive conditions. The current phase delay through the delay line is less than 90 degrees, (90 degrees − 16.3 degrees, or 73.7 degrees, the difference between the input and output angle). Observe that the input voltage amplitudes and phases are not alike at the input ends of the coaxial lines from the two elements. But these two terminals are normally connected together; clearly two different voltages can't coexist here. Since the difference is fairly small, the actual drive conditions that result if connected anyway will be acceptable, though the F/B ratio will diminish. The choice of 1/4-wavelength element feeders just happened to provide this fair agreement. I estimate the actual phase displacement between elements to be about 115 degrees and the current amplitude ratio about 1.15. The 1/4-wavelength delay line didn't produce a 90-degree delay and the delays in the two equal length feeder lines were unequal; these are all quite different results from what is often assumed to occur.

Some writers have assured us we can use any

length feeders as long as lengths are kept equal. Let's see how we fare following this advice using *3/8-wavelength 50-ohm feeders* for the same array:

Element 1	**Element 2**
$Z_1 = 21.4 - j15$	$Z_2 = 51.4 + j15$
$I_1 = 1 \angle 0^\circ$	$I_2 = 1 \angle -90^\circ$
$E_1 = 26.13 \quad \angle -35.03^\circ$	$E_2 = 53.54 \quad \angle -73.73^\circ$
135° Feeder	**135° Feeder**
$Z_1 = 63.58 - j53.98$	$Z_2 = 37.43 + j2.67$
$I_1 = 0.58 \quad \angle 148.56^\circ$	$I_2 = 1.17 \quad \angle 51.66^\circ$
$E_1 = 48.39 \quad \angle 108.22^\circ$	$E_2 = 43.97 \quad \angle 55.75^\circ$
	1/4 wavelength delay line
	$Z_2 = 66.46 - j4.75$
	$I_2 = 0.88 \quad \angle 145.75^\circ$
	$E_2 = 58.60 \quad \angle 141.66^\circ$

Note: all impedances, voltages, and currents are input conditions — that is, looking towards the load.

Using 3/8-wavelength feeders, the input voltages required to be applied to each chain are very different. If these terminals are tied together the drive conditions to the elements will be far from intended. Conclusion: element feeders are an integral part of the feed network for a phased array; their circuit characteristics must be taken into account.

designing for optimum drive

While it is possible to solve for the undesirable drive conditions that would result from making this connection, why bother? It is better to start with the correct design. While doing so, suppose a 1:1 SWR match to a 50-ohm array feedline is included. This would require that the paralleled input impedances of the networks from each element be 50 ohms pure resistance. Assuming lossless conditions, we can go back to the resistive components of the element driving-point impedances for this determination. These are 21.4 and 51.4 ohms, respectively. At 1 ampere to each element the total drive power is the sum of the I^2R inputs, or 72.8 watts. Using the relationship $E^2/R = W$, and substituting 50 ohms (the characteristic impedance of the array feed-line) for R:

$$E^2 = 50(72.8), \text{ or } E = \sqrt{3640} = 60.33 \text{ volts}$$

Having established the array feedline voltage amplitude for this drive power, we can calculate the required resistive inputs for each element's network. Rearranging, $R = E^2/W$:

$$R_1 = 3640/21.4 = 170.09 \text{ ohms for network, No. 1 input}$$

$$R_2 = 3640/51.4 = 70.82 \text{ ohms for network, No. 2 input}$$

As a check on calculations it is useful to do the parallel conductance calculation:

$$1/R_1 + 1/R_2 + \cdots + 1/R_n = 1/Z_0 \qquad (2)$$

Starting again at the end of the 3/8-wavelength feeder to element 1, one possibility is an L-match network, transforming directly to 170.09 ohms pure resistance. L-match circuit component calculations involve a square root extraction, guaranteeing at least two solutions. (Under certain circumstances, there may be four solutions.) While all solutions produce the intended transformation, they do so with differing phase displacements, with at least one of those displacements being a leading phase. Remembering that element 2 is starting 90 degrees behind the first, fewer stages in the network usually result if a leading phase L-match is chosen for element 1.

Shown in **table 2**, beginning with the driving-point impedances and working forward to the common

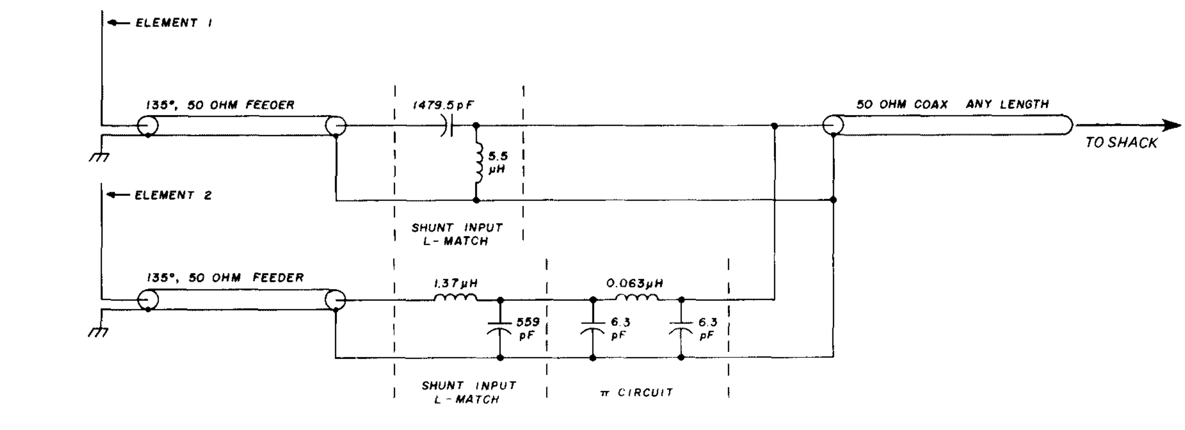

fig. 2. Matching network for a 2-element array using 3/8 λ coaxial feeders at a 3.8 MHz design frequency.

connection of the array, are the input parameters of each circuit. **Fig. 2** shows the schematic and component values at the design frequency of 3.8 MHz.

At the inputs of each network chain, E_1 and E_2 are equal in amplitude and phase; the two inputs may be connected together without disturbing drive conditions. Their paralleled resistive inputs represent a 50-ohm resistive load, as designed. The π coaxial-equivalent network was added to the element 2 chain only to show how this type of network circuit is used to match the voltage phase at the common connection of the array network. In this example the agreement that happened to be achieved at the input to the 2 element shunt L-match is sufficient; the π network can be omitted.

4-terminal networks. The design procedure for producing exact matching at the required array conditions has been demonstrated. Before proceeding with other examples, the design equations for these circuits are presented. All network circuit components are reactances and assumed to be lossless. Subscript a denotes the series load termination components, $R_a + jX_a$, instead of the more commonly used $R_L + jX_L$, to avoid any confusion with jX_L, as an inductive reactance.

L-match circuit. This circuit can take two forms (see **fig. 1A** and **1B**), termed *Shunt Input* and *Series Input* L-matches. Though this circuit consists of only two components, its analysis is relatively complex. The calculations for either form include a square root extraction, resulting in two possible sets of components for any desired impedance transformation. Either set works; one set often has a leading phase angle, while the other may lag. The absolute value of the angles are not necessarily equal nor always of opposite sign! The component set may have the same reactance sign; that is, both may be inductive (+) or both capacitive (−). The circuit is sometimes referred to as an L-C match, but it could also be an L-L or C-C match. A more apt description is L-match, taken from the similarity of its schematic representation to the letter "L".

Shunt input L-match. The series arm component X_2, must be calculated first, since its value is used in the calculation for the second component:

$$X_2 = -X_a \pm \sqrt{R_a(R_{in} - R_a)} \text{ ohms} \quad \textbf{(3)}$$

$$X_1 = -\left[\frac{R_a^2 + (X_2 + X_a)^2}{X_2 + X_a}\right] \text{ ohms} \quad \textbf{(4)}$$

where R_a and X_a are the series equivalents of the load termination and
R_{in} is the desired input (pure) resistance.

Close attention must be paid to signs. A positive result indicates an inductance, while a negative sign is a capacitance.

Series input L-match. X_2 must be calculated first. Note that X_2 is the shunt arm of this circuit, however.

$$X_2 = \frac{-R_{in}X_a \pm \sqrt{R_{in}R_a(R_a^2 + X_a^2 - R_{in}R_a)}}{R_{in} - R_a} \text{ ohms} \quad \textbf{(5)}$$

$$X_1 = \frac{-X_2[R_a^2 + X_a(X_2 + X_a)]}{R_a^2 + (X_2 + X_a)^2} \text{ ohms} \quad \textbf{(6)}$$

Which form should be used? Usually, the shunt input L-match is the only form possible if R_{in} is equal to or greater than R_a. Besides, the arithmetic is easier! The series input L-match is used when R_{in} is less than R_a. There is a set of circumstances, however, in which the series form can be used even if R_{in} is greater than R_a. Inspection of the equation for the series form calculation of X_2 will show this case when R_{in} is greater than R_a and when $(R_a^2 + X_a^2 - R_{in}R_a)$ is equal to or greater than zero. Four solutions, two series and two shunt L-matches, then result. These additional options, if available, are often useful, allowing a smaller phase displacement or more (physically) realizable set of components as a result.

π coaxial-equivalent circuit (fig. 1C). The π circuit and the shunt input L-match will be found to be the most frequently used circuits for vertical phased array feed networks. The type used here is set up as a reversible network — that is, the input and output can be interchanged without affecting operation, just as with coaxial cable. Reactances X_1 and X_3 are always equal and if capacitive, then X_2 is inductive. At the design frequency this particular configuration shows the same properties as coax. As the frequency is varied, the phase displacement starts differing from that obtained with coax, the difference being larger the greater the equivalent "length" of the circuit. If, instead, multiple sections, each an equal increment of the total phase displacement, are cascaded, the combined network approaches coaxial cable characteristics. This should be expected since the equivalent circuit of coax is a series of infinitesimally small π sections. The design equations are relatively simple:

$$X_2 = Z_0 \sin\theta \text{ ohms} \quad \textbf{(7)}$$

$$X_1 = X_3 = -\frac{Z_0 \sin\theta}{1 - \cos\theta} \text{ ohms} \quad \textbf{(8)}$$

where Z_0 is the required characteristic impedance
θ is the electrical length in degrees

A positive sign indicates an inductance while a negative sign indicates capacitance.

If a *leading* phase, say 30 degrees, is desired, this would be equivalent to 330 degrees in electrical

length if coaxial cable were used. Substituting 330 degrees in these equations causes X_2 to be negative and X_1 and X_3 to be positive; the appropriate capacitance and inductances can then be calculated from the relations:

$$C = 1/\omega X \text{ and } L = X/\omega \quad \textbf{(9)}$$

$$\text{where } \omega = 2\pi f, f = \text{frequency in Hz}$$

Half-wave section (180 degree electrical length) π circuits are taboo, since the calculated circuit values are physically unrealizable. At the least, two separate 90-degree sections are suggested to achieve this "electrical length."

Tee coaxial-equivalent circuit (fig. 1D). This circuit is used in the same applications as the π circuit. Alternated with π networks in equal increments of electrical length, network characteristics can be made to equal or exceed coax (assuming coax of the same characteristic impedance is available for comparison). For applications requiring a leading phase displacement only one inductance (for the shunt arm) is necessary, sometimes simplifying construction. The design equations are:

$$X_2 = -Z_0/\sin\theta \text{ ohms} \quad \textbf{(10)}$$

$$X_1 = X_3 = \frac{Z_0(1 - \cos\theta)}{\sin\theta} \text{ ohms} \quad \textbf{(11)}$$

where Z_0 = required characteristic impedance
θ = electrical length required in degrees

As with the π network, a positive sign indicates inductive reactance and a negative sign, capacitive reactance. Also, 180 degree sections cannot be physically realized and require at least a 2-section cascaded network to achieve that displacement.

Series impedance circuit (fig. 1E). This circuit is used when R_{in} is equal to R_a and the load has a reactance X_a. The series matching impedance is simply the reactance of the opposite sign.

$$X = -X_a \text{ ohms} \quad \textbf{(12)}$$

Parallel impedance circuits (fig. 1F).

$$X = -\left[X_a + \frac{R_a^2}{X_a}\right] \text{ ohms} \quad \textbf{(13)}$$

The parallel matching reactance has the opposite sign of the parallel equivalent reactance of the load. The series and parallel circuits can be thought of as a shunt input L-match — with one of its circuit branches either equal to infinity or zero impedance, respectively.
These conditions occur when $R_{in} = R_a$ or

$$X_a = \sqrt{R_a(R_{in} - R_a)}$$

Either circuit should be considered, particularly when the load has a relatively large reactance compared to its resistive component. The circuit is simple, and cascaded with a following L-match circuit, results in a broader bandwidth network.

design limitation and other considerstions

Some design hints may be helpful to understanding the use of these circuits:

1. The L-match circuits first require selection of the input resistance wanted, transforming from any output impedance. Phase displacement, however, cannot be pre-defined, though the direction, lead or lag, may be chosen.

2. Single L-match impedance transformation ratios exceeding 5-to-1 should be avoided. Above that ratio, expect to see increased frequency sensitivity and resultant reduction in bandwidth. For high ratios, consider transforming in step increments of resistance using several L-matches or combinations of L-match and π or tee circuits (the latter as 1/4-wavelength transformers).

3. In this particular application, π and tee circuits are always designed for pure resistance terminations. These circuits are designed to act as a 1/4-wavelength transformer, or as a specific coaxial-equivalent length, leading or lagging, of transmission line. Choose any characteristic impedance, but keep in mind that large (more than ± 90 degree) increments of angular displacement, especially at high impedances, reduce bandwidth.

4. Cascaded circuits may each have a capacitor at their common connection points, which are then in parallel. For example, see **fig. 3** showing a 567 pF and 392 pF capacitor at a common connection point. The two values may be added and a single capacitor placed at that junction. However, until the network has been tested, it is useful to keep the circuits independent for separate adjustment.

designing networks for multi-element arrays

Armed with the design equations for simple 4-terminal networks, we can now examine feed networks for arrays consisting of several elements. If the array is one requiring a phase angle multiple, for example, 0, 90, and 180 degrees, or 0, 100 and 200 degrees, and all feedlines are equal in length, the simplest network may result if the middle element is treated as if it were the reference element of the array. The respective networks for the array end elements are designed to lead and to lag the middle element. Then neither has to be designed to span a large angular displacement, and fewer stages result.

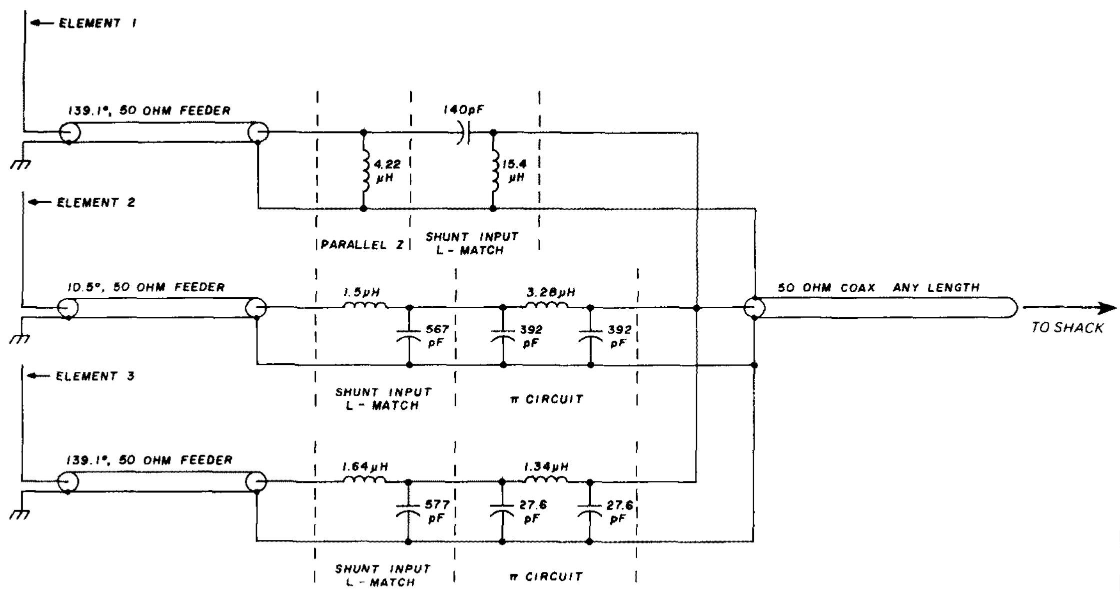

fig. 3. Matching network for a 3-element in-line array at a 3.8 MHz design frequency.

3-element in-line array. This array has a particularly deep F/B ratio extending over a wide azimuthal sector. We should be especially interested in taking advantage of this capability. Since the middle element has the same drive-point impedance regardless of array direction, there is no need to make its feeder equal in length to other feeders. Assuming the directional switch is located five feet from the middle element, equal length end element feeders are brought to the center area. At 3.8 MHz, using 0.66 velocity factor coax, these are 66 feet (139.1 degrees) and for the center element, 5 feet (10.5 degrees) with a Z_0 of 50 ohms. Assuming an array of 3 resonant 1/4-wavelength elements, spaced a quarter-wavelength apart, with current amplitude ratios of 1,2,1 and phase relationships of 0, −90, and −180 degrees, respectively, the driving point impedances are $Z_1 = 15.4 - j17$, $Z_2 = 36.2 + j0$ and $Z_3 = 75.4 + j43$. (Part 3 showed these values incorrectly).[5] As was done with the 2-element array example, the feed network is matched to the 50-ohm array feedline. The sum of the I^2R input power terms, assuming 1 ampere to the first and third elements and 2 amperes to the middle element, is 235.6 watts. Using the $E^2/R = W$ relationship, this establishes an amplitude of 108.54 volts at the array feedline connection. At that point the input impedances for each element's network are the pure resistances:

$$Z_1 = 764.94 + j0$$
$$Z_2 = 81.25 + j0$$
$$Z_3 = 156.23 + j0$$

The sequence of input parameters at each junction of the networks is shown in **table 3**.

The resulting network is shown in **fig. 3**. Illustrated in this example is the application of the parallel circuit and the use of leading and lagging phase L-match circuits. Here, element 1 is used as the reference element of the feed network. A parallel impedance circuit is used to transform the impedance seen at the input end of the feeder to a pure resistance. This is then transformed to the pure resistance required for the chain with a shunt input L-match chosen to produce a leading phase change. The resulting input voltage then becomes the objective for the other two network chains.

Triangular array. The triangular array feed network demonstrates still another technique for simplifying a feed network. Since elements 2 and 3 operate at identical conditions, the inputs of their transmission line feeders may be paralleled and fed from a common network. **Fig. 4** shows the two feeders connected to a shunt input L-match, and being transformed directly to a resistive input. This is then cascaded with a tee circuit having a sufficiently leading phase displacement to equal the voltage amplitude and phase of the element 1 network. The array termination is designed to match a 50-ohm transmission line. Part 3 incorrectly showed the driving-point impedance of element 1. The correct impedance is $Z_1 = 20.4 - j10$. **Table 4** shows the sequence of input parameters at each network junction.

4-square array. The 4-square array obviously requires a more complicated drive network. The look-

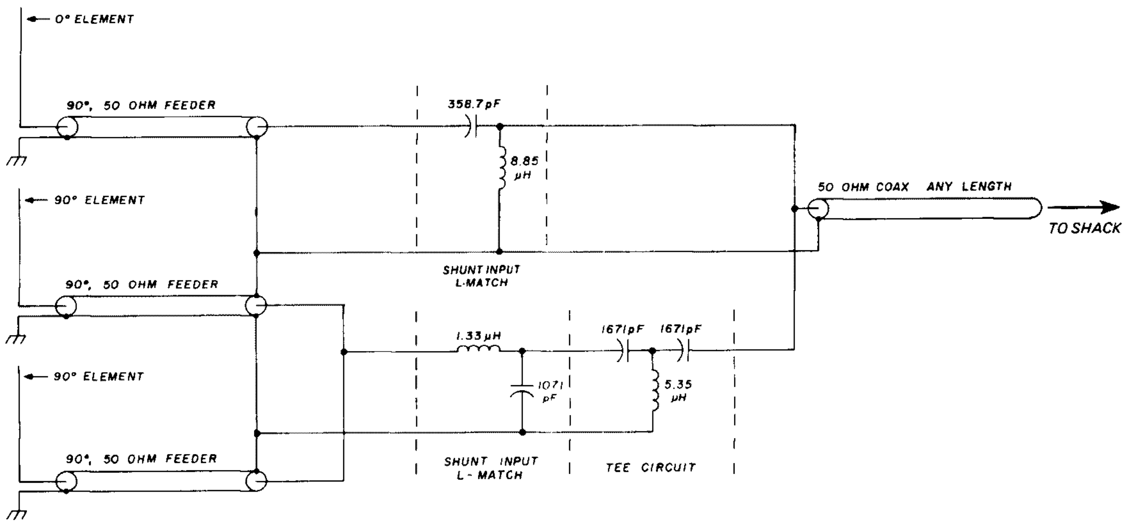

fig. 4. Matching network for a triangular array at a 3.8 MHz design frequency.

alike middle elements present the opportunity to connect their feedlines in parallel, simplifying the design somewhat. The 4-square element driving-point impedances are highly reactive, making any drive network more frequency dependent. There is the further question of the directions the driving-point impedances take as frequency is changed from design center. The relatively small amount of measurements I have taken to examine this question indicate a not unexpected similarity to Yagis. Array performance falls apart more rapidly on the low-frequency side of design center than on the high side. Whether a drive

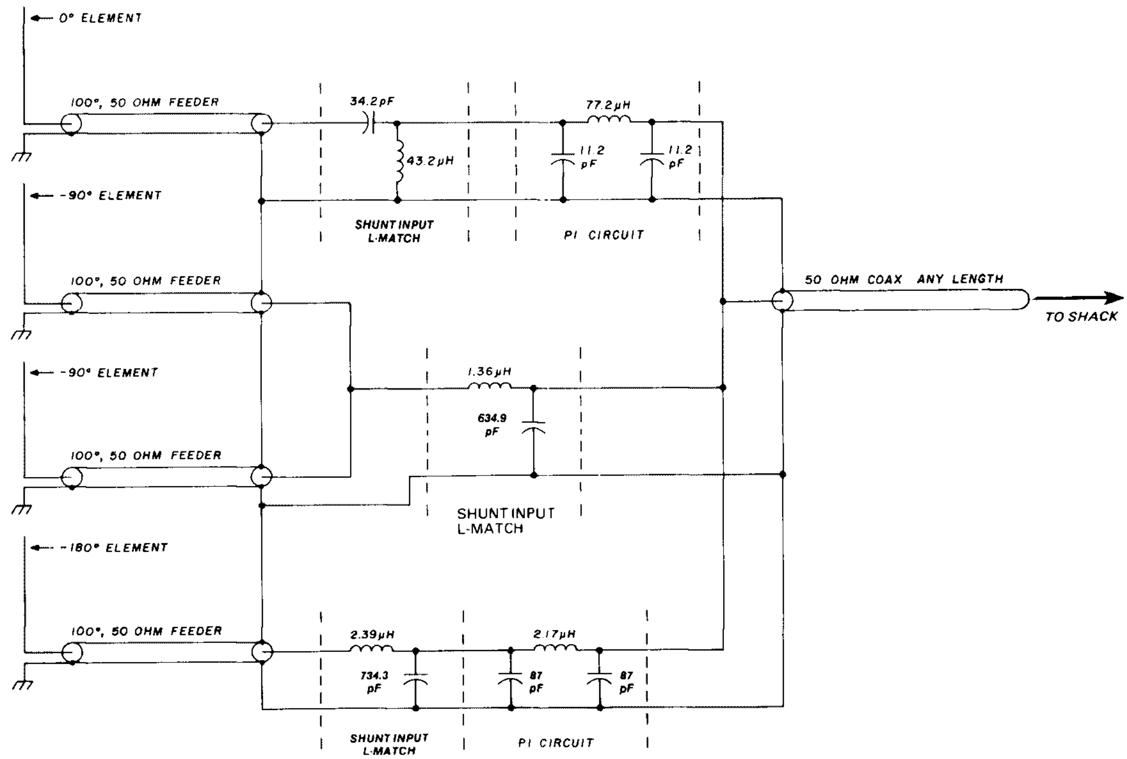

fig. 5. Matching network for a 4-square array at a 3.8 design frequency.

table 1. Comparison of impedance, voltage and current phase, and SWR variations with frequency, of 90 degree length of coax and a cascaded 45 degree π circuit and 45 degree tee circuit, both acting as a 50-ohm characteristic impedance 1/4-wavelength transformer. Load termination is 75 ohms pure resistance; the center design frequency is 3.75 MHz.

frequency MHz		coax	π and tee
3.55	Z_{in}	33.46 − j2.33	33.45 − j1.86
	I_o/I_i	0.67 $\angle -86.8°$	0.67 $\angle -86.33°$
	E_o/E_i	1.49 $\angle -82.82°$	1.50 $\angle -83.15°$
	SWR	1.5	1.498
3.65	Z_{in}	33.37 − j1.16	33.36 − j0.91
	I_o/I_i	0.67 $\angle -88.4°$	0.67 $\angle -88.15°$
	E_o/E_i	1.5 $\angle -86.4°$	1.5 $\angle -86.58°$
	SWR	1.5	1.5
3.75	Z_{in}	33.33 + j0	33.33 + j0
	I_o/I_i	0.67 $\angle -90°$	0.67 $\angle -90°$
	E_o/E_i	1.5 $\angle -90°$	1.5 $\angle -90°$
	SWR	1.5	1.5
3.85	Z_{in}	33.37 + j1.16	33.36 + j0.88
	I_o/I_i	0.67 $\angle -91.6°$	0.67 $\angle -91.88°$
	E_oE_i	1.5 $\angle -93.6°$	1.5 $\angle -93.39°$
	SWR	1.5	1.5
3.95	Z_{in}	33.46 + j2.33	33.45 + j1.72
	I_o/I_i	0.67 $\angle -93.2°$	0.67 $\angle -93.79°$
	E_o/E_i	1.49 $\angle -97.18°$	1.5 $\angle -96.74°$
	SWR	1.5	1.498

Note: I_i, E_i equal input current and voltage, respectively. I_o, E_o equal output current and voltage, respectively.

network can be designed to reduce this tendency is a question. Perhaps the best alternative is to set the design center frequency on the low end of the intended operating range, recognizing that the optimum F/B bandwidth is a narrow frequency band of about 2 to 3 percent.

Using the driving-point impedances from Part 3 for a 4-square consisting of 1/4-wavelength resonant elements, spaced a quarter-wavelength apart, and phased 0, −90, −90, −180 degrees with current amplitude ratios 1,1,1,1, respectively, input parameter sequences of a suggested drive network are

table 5. Network input parameters for a *4-square array*.

element 1	element 2	element 3	element 4
Z_1 3.4 − j12	Z_2 39.4 − j17.5	Z_3 39.4 − j17.5	Z_4 63.4 + j47.5
I_1 1 $\angle 0°$	I_2 1 $\angle -90°$	I_3 1 $\angle -90°$	I_4 1 $\angle -180°$
E_1 12.47 $\angle -74.18°$	E_2 43.11 $\angle -113.95°$	E_3 43.11 $\angle -113.95°$	E_4 79.22 $\angle -143.16°$
100° coax	**100° coax**	**100° coax**	**100° coax**
Z_1 403.97 + j387.1	Z_2 62.39 + j22.57	Z_3 62.39 + j22.57	Z_4 22.73 − j11.37
I_1 0.09 $\angle 46.88°$	I_2 0.79 $\angle -12.43°$	I_3 0.79 $\angle -12.43°$	I_4 1.67 $\angle -74.97°$
E_1 51.33 $\angle 90.66°$	E_2 52.73 $\angle 7.46°$	E_3 52.73 $\angle 7.46°$	E_4 42.44 $\angle -74.97°$
	elements 2 & 3 paralleled		
L-match	**L-match**		**L-match**
Z_1 2141 + j0	Z_2, Z_3 92.39 + j0		Z_4 114.83 + j0
I_1 0.04 $\angle -17.37°$	I_2, I_3 0.92 $\angle 42.04°$		I_4 0.74 $\angle 15.2°$
E_1 85.32 $\angle -17.37°$	E_2, E_3 85.32 $\angle 42.04°$		E_4 85.32 $\angle 15.2°$
π circuit			**π circuit**
Z_1 2141 + j0			Z_4 114.83 + j0
I_1 0.04 $\angle 42.04°$			I_4 0.74 $\angle 42.04°$
E_1 85.32 $\angle 42.04°$			E_4 85.32 $\angle 42.04°$

table 2. Network input parameters for a *2-element array*.

element 1		element 2	
Z_1	21.4 − j15	Z_2	51.4 + j15
I_1	$1 \angle 0^\circ$	I_2	$1 \angle -90^\circ$
E_1	$26.13 \angle -35.03^\circ$	E_2	$53.54 \angle -73.73^\circ$
135° coax feeder		**135° coax feeder**	
Z_1	63.58 − j53.98	Z_2	37.43 + j2.67
I_1	$0.58 \angle 148.56^\circ$	I_2	$1.77 \angle 51.66^\circ$
E_1	$48.39 \angle 108.22^\circ$	E_2	$43.97 \angle 55.75^\circ$
shunt input L-match		**shunt input L-match**	
Z_1	170.09 + j0	Z_2	70.82 + j0
I_1	$0.36 \angle 96.25^\circ$	I_2	$0.85 \angle 95.03^\circ$
E_1	$60.33 \angle 96.25^\circ$	E_2	$60.33 \angle 95.03^\circ$
		π circuit	
		Z_2	70.82 + j0
		I_2	$0.85 \angle 96.25^\circ$
		E_2	$60.33 \angle 96.25^\circ$

table 3. Network input parameters for a *3-element in-line array*.

element 1		element 2		element 3	
Z_1	15.4 − j17	Z_2	36.2 + j0	Z_3	75.4 + j43
I_1	$1 \angle 0^\circ$	I_2	$2 \angle -90^\circ$	I_3	$1 \angle -180^\circ$
E_1	$22.94 \angle -47.83^\circ$	E_2	$72.4 \angle -90^\circ$	E_3	$86.80 \angle -150.3^\circ$
139.1° Coax		**10.5° Coax**		**139.1° Coax**	
Z_1	47.42 − j67.6	Z_2	36.71 + j4.35	Z_3	27.77 + j20.68
I_1	$0.57 \angle 159.27^\circ$	I_2	$1.98 + \angle -82.33^\circ$	I_3	$1.65 \angle -36.83^\circ$
E_1	$47.06 \angle 104.32^\circ$	E_2	$73.49 \angle -75.59^\circ$	E_3	$56.98 \angle -0.25^\circ$
parallel Z		**shunt input L**		**shunt input L**	
Z_1	143.19 + j0	Z_2	81.35 + j0	Z_3	156.23 + j0
I_1	$0.33 \angle 104.32^\circ$	I_2	$1.33 \angle -34.59^\circ$	I_3	$0.69 \angle 28.24^\circ$
E_1	$47.06 \angle 104.32^\circ$	E_2	$108.54 \angle -34.59^\circ$	E_3	$108.54 \angle 28.24^\circ$
shunt input L		**π circuit**		**π circuit**	
Z_1	764.94 + j0	Z_2	81.35 − j0	Z_3	156.23 + j0
I_1	$0.14 \angle 40.01^\circ$	I_2	$1.33 \angle 40.01^\circ$	I_3	$0.69 \angle 40.01^\circ$
E_1	$108.54 \angle 40.01^\circ$	E_2	$108.54 \angle 40.01^\circ$	E_3	$108.54 \angle 40.01^\circ$

given in **table 5** with the schematics shown in **fig. 5**.

Reaching the center area of a 4-Square array whose elements are spaced 1/4 wavelength with 1/4 wavelength feeders is a problem. Coax with a velocity factor greater than 0.71 is required. Unfortunately, most foam dielectric cables have velocity factors around 0.71, allowing little or no slack for placement of and connections to a directional switch and the feed network. Since a lumped-constant network imposes no restriction on element feeder lengths, I have chosen 100° for these feeders (at 3.8 MHz, 47.45 feet), which is more than sufficient.

The input ends of the feeders to the two middle elements are paralleled and a shunt input L-match is used to transform this combined load directly to the desired pure resistance at the common connection to the array. The input voltage for this chain is the design objective for the networks feeding the remaining elements. Consequently, after transforming the other elements to their required pure resistances (equalling 50 ohms when all networks are paralleled), coax-equivalent π circuits are used to match the voltage phase required at the common connection of the array.

table 4. Network input parameters for a *triangular array*.

element 1		element 2		element 3	
Z_1	20.4 − j10	Z_2	78.4 + j4	Z_3	78.4 + j4
I_1	$1 \angle 0^\circ$	I_2	$0.5 \angle -90^\circ$	I_3	$0.5 \angle -90^\circ$
E_1	$22.72 \angle -26.11^\circ$	E_2	$39.25 \angle -87.08^\circ$	E_3	$39.25 \angle -87.08^\circ$
90° Coax		**90° Coax**		**90° Coax**	
Z_1	98.81 + j48.43	Z_2	31.81 − j1.62	Z_3	31.81 − j1.62
I_1	$0.45 \angle 63.89^\circ$	I_2	$0.79 \angle 2.92^\circ$	I_3	$0.79 \angle 2.92^\circ$
E_1	$50 \angle 90^\circ$	E_2	$25 \angle 0^\circ$	E_3	$25 \angle 0^\circ$
shunt input L		**elements 2 & 3 paralleled shunt input L**			
Z_1	146.08 + j0	Z_2, Z_3	76.02 + j0		
I_1	$0.37 \angle 29.22^\circ$	I_2, I_3	$0.72 \angle 65.70^\circ$		
E_1	$54.59 \angle 29.22^\circ$	E_2, E_3	$54.59 \angle 65.70^\circ$		
		tee circuit			
		Z_2, Z_3	76.02 + j0		
		I_2, I_3	$0.72 \angle 29.22^\circ$		
		E_2, E_3	$54.59 \angle 29.22^\circ$		

coming soon

In a forthcoming article in this series I shall cover in detail a method of calculating simple 4-terminal networks based on matrix algebra. As with many mathematical procedures, application of the procedures to the solution of problems doesn't necessarily depend upon a complete understanding of the underlying theory. It is a tool simplifying what is otherwise a tedious process. The calculation procedures are structured (differing equations for matrix values), making it ideal for programmable calculators or small computers. For the same reason, the procedure easily lends itself to chained calculations, allowing fast analysis of a cascade of networks.

reference

1. Forrest Gehrke, K2BT, "Vertical Phased Arrays, Part 1," *ham radio*, May, 1983, page 18.
2. Forrest Gehrke, K2BT, "Vertical Phased Arrays, Part 2," *ham radio*, June, 1983, page 24.
3. Forrest Gehrke, K2BT, "Vertical Phased Arrays, Part 3," *ham radio*, July, 1983, page 26.
4. Roy Lewallen, W7EL, private communication. Contact K2BT for further information.
5. See "Short Circuits," *ham radio*, October, 1983.

ham radio

a one-weekend, 2-meter amplifier

30 watts from an easy-to-build project using low-cost components

After years of operating on two-meter FM with a tube-type radio, I decided it was finally time to replace it with something less demanding on the car's electrical system. My purchase of a synthesized hand-held transceiver spurred me to develop an amplifier which would provide a mobile system with the same capabilities as the tube rig, but with lower power consumption. That amplifier is described here.

evolution of the amplifier

I've never seen an amplifier design that is quite what I needed to meet these design requirements: 1-watt drive for full output (30 to 40 watts), low cost, ease of construction, and long-term reliability. I decided to build one based on a two-transistor circuit.

The 2N6080 and MRF238 transistors were readily available at low cost. This particular combination provided an excess of drive power between stages; however, this helps make the amplifier wideband, and ensures continued operation near rated output under low-battery conditions. The power consumption consequently is slightly higher than necessary, but still much less than that of an equivalent tube-type radio.

To make the amplifier easy to duplicate and tune, I decided to use printed-line inductors instead of discrete coils. The next step was to calculate the impedance-matching networks. Motorola's excellent application notes[1] were used to do this. Once calculated, the networks were verified using "The Electronic Breadboard."[2] Data for the inductors was obtained from references 3 and 4. Once built, the matching networks were modified empirically.

The resulting circuit is shown in **fig. 1**. The input-matching network consists of C_1, C_2, C_3, C_{13}, and L_1. The interstage network uses C_4, C_5, and L_2, and the output network is made up of L_3, C_6, C_7, C_8 and C_{14}. Power-supply decoupling is very important and is achieved by C_9-C_{12}, RFC2, RFC4, and feed-through capacitor C_{15}.

The 1N540 diode serves two functions. It protects the amplifier by clipping inductive spikes from things like starter motors. It also protects the amplifier against reverse polarity of the power supply, by causing the line fuse to blow.

Once the amplifier was constructed and working, some means of transmit/receive switching was needed. PIN diodes are state-of-the-art and small, but not readily available. The alternative was to modify an open-frame relay, as described in reference 5. The relay circuit is actuated by RF from the exciter. Its driver circuit is shown in **fig. 2**.

construction

Once the design is completed, the circuit board must be constructed. I used a 9-1/4 × 2-3/4 inch (23.5 × 7 cm) piece of G-10 board with copper on both sides. Orr[6] highly recommends short, direct paths, and this philosophy was followed. The inductor and RF-transistor connecting pads must be laid out first. It is important to realize that these are printed inductors and not transmission lines; therefore, the copper material on the reverse side of the board must be removed under the inductors. If this is not done, the amplifier will not work.

Although straight-line inductances are shown in **fig. 3**, this is not mandatory. The important parameters are average path length and the path width. Next, lay out the power-supply line — remember it carries several amperes and must not have any appreciable resistance. Finally, allocate an area for the modified T/R relay and its driver circuit, and lay this out. The layout that I used is shown in **fig. 3**.

This layout is simple enough to reproduce directly on the board, using ordinary electrical tape and trimming it to size with a sharp blade. After the layout is completed and verified, etch the board using standard techniques. (I used ferric chloride.)

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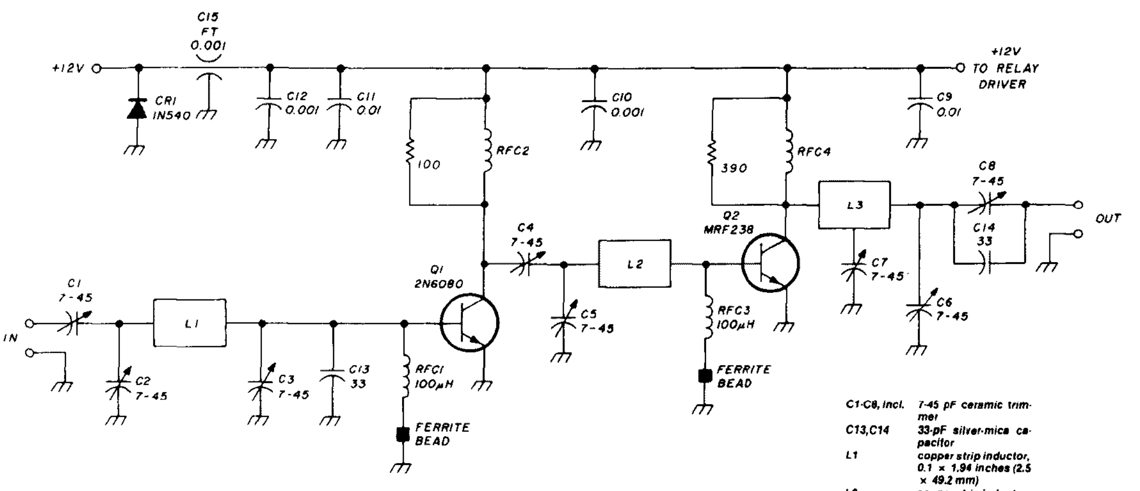

Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms.
k = 1,000 M = 1,000,000

C1-C8, incl.	7-45 pF ceramic trimmer
C13,C14	33-pF silver-mica capacitor
L1	copper strip inductor, 0.1 × 1.94 inches (2.5 × 49.2 mm)
L2	copper strip inductor, 0.13 × 0.72 inch (3.3 × 18.3 mm)
L3	copper strip inductor, 0.13 × 2.22 inches (3.3 × 56.4 mm)
RFC2	9 turns No. 18 enameled, wound on 100-ohm, 1-watt resistor
RFC4	5 turns No. 18 enameled, wound on 390-ohm, 2-watt resistor

fig. 1. Schematic diagram of the 30-watt, two-meter amplifier.

After the board is made and drilled, mount it to the heatsink using 1/8 inch (0.32 cm) spacers 5/16 inch (0.79 cm) in diameter. Mount the RF transistors to the heatsink, using a thermal compound to ensure good heat transfer. If a torque wrench is available, torque the hardware to six inch pounds (0.68 Ntm). Otherwise, tighten it until a moderate resistance is felt, using a wrench no longer than three inches. Make sure you hold the stud to prevent it from moving. Avoid overtorquing the stud on the transistors or they could be damaged. Also, be sure that the top and bottom ground planes are connected together in at least six places.

After the transistors are mounted to the heatsink, solder them to the PC board. The transistors are mounted before soldering to avoid excessive (damaging) stresses on the leads during mounting to the heatsink.

Next, mount the capacitors and chokes, keeping the lead lengths as short as possible. Follow the parts layout in **fig. 4**. Mount the relay, using either metal hardware or epoxy glue. Wire the input and output coax lines to the relay, and then install the relay-driver components. The input and output coax connectors can be mounted directly on the PC board, or made part of the cabinet and connected by short lengths of coax. This completes the actual construction.

Carefully check the wiring against the circuit diagram. Pay close attention to transistor orientation and placement of the variable capacitors. The amplifier is now ready for a "smoke test."

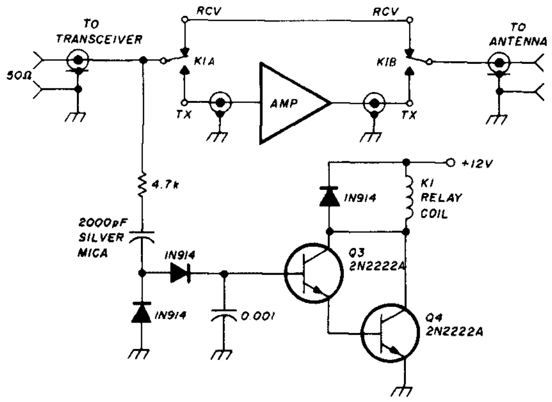

fig. 2. Transmit/receive relay control circuit.

testing

Connect the amplifier to a 12-volt source through a fuse of 10 amperes, a 50-ohm load, and a driver. Apply power and monitor the idling current — it should be low (a few mA). With some means of measuring output to the load, apply drive to the amplifier. The relay should switch to the transmit position. If it doesn't, check the relay-control circuit and try again.

Once the relay transfers, you should see some RF output. If not, adjust the input tuning until you see some low value of output. Then peak the output network for maximum output. Next, peak the interstage and input networks for maximum output. Repeat the entire procedure until no further increase in output

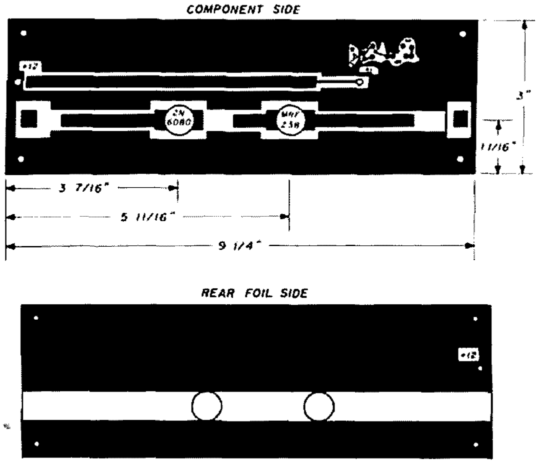

fig. 3. (A) Layout of the top (component) side of the printed circuit board; (B) Layout of the bottom side of the printed circuit board.

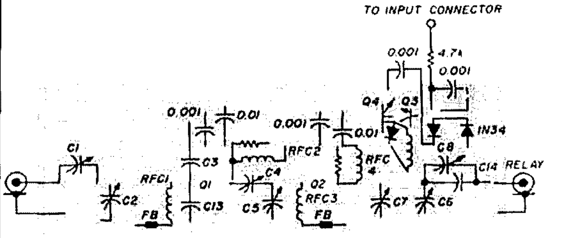

fig. 4. Parts placement on the component side of the printed circuit board.

can be seen. The output should be between 30 and 40 watts, depending on drive level and power-supply voltage. The input SWR should be very close to 1:1. Measure the power supply current, which should be between 4 and 6 amps. Keep in mind that the duty cycle should be kept low until the tuning has been completed. Pay close attention to transistor temperatures — they will get warm, and excessive temperature can result in catastrophic failure.

final assembly

The enclosure is up to you. I soldered together double-sided circuit board to make a simple, cheap, and RF-tight enclosure. After the amplifier is placed in the enclosure, recheck the output — there should be little change.

Amateurs often ignore a basic fact of heat transfer and fluid flow: hot air rises. The heatsink needs to be mounted in such a way that the hot air can flow unimpeded. This means that the fins must be vertical. I have seen far too many installations where heatsink fins are horizontal, reducing efficiency. The obvious benefit of doing this right is increased reliability and long life.

results

I installed the amplifier in my car and it has been working well for several months. Output power and gain as a function of frequency are shown in **fig. 5.** Though not perfect, the bandwidth is certainly acceptable for use on the two-meter band. During several months of operation, no degradation in performance or change in tuning has been noted. This attests to the long-term stability and reliability of the amplifier.

One aspect all too often neglected by Amateurs is that of spurious outputs. This amplifier was checked on a spectrum analyzer with gratifying results. There were no measurable outputs other than the one desired. As close as could be measured, the output was a faithful reproduction of the input. When modulated with a single tone, the TR-2400-and-amplifier combination exhibited textbook response, clean and symmetrical, with no hint of instability or distortion. The insertion loss in the receive mode is negligible.

The total time needed to build and test this amplifier was less than one weekend, including fabricating the circuit board. The final product was a reliable 30-watt amplifier capable of being driven to full output by a 1-watt hand-held transceiver.

parts

One of my pet peeves is people who describe equipment that is very nice to build, but uses hard-to-find, or outrageously expensive, custom-manufactured components. This amplifier uses readily available components, and I will cite my sources for non-

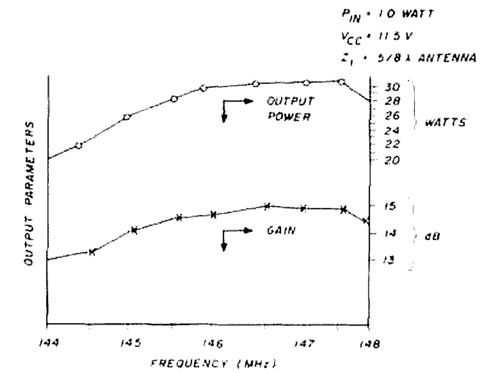

fig. 5. Plot of output power and amplifier gain as a function of frequency. Note that the response is useful over much of the two-meter band.

junk-box parts. The transistors came from Semiconductor Surplus. I obtained the circuit board and ceramic trimmers from a local surplus outlet, but these are seen at every hamfest and in the advertising pages of *ham radio*. The chokes came from the junk box and are not critical. The ferrite beads are available from Amidon and others, and also at hamfests. I have modified several types of junk-box relays, and all worked well in this circuit.

You can make some substitutions. The chokes are not critical; just be sure to use ferrite beads to lower the Q and prevent low-frequency oscillations. I used ceramic trimmer capacitors, although compression or piston types would probably also work. Whatever type is used, it must be capable of withstanding the rf currents involved. This is especially important in the output circuitry. If you have access to a good capacitance meter, you can build the amplifier using various capacitors, and then substitute fixed-value chip capacitors, such as those made by Unelco. This will provide maximum reliability.

The most critical part is the heatsink — if it is not adequate, the transistors will not survive. I used one 6 × 3-3/4 inches (15.24 × 9.83 cm). If you have a bigger one, so much the better. The more area available for heat transfer, the cooler the transistors will run.

To ensure adequate cooling in the hot Florida sun, I placed a fan on the heatsink to help keep temperatures low. This may not be necessary, but is good insurance to help the amplifier last. To date there have been no thermal problems, even with temperatures inside the car higher than 100 degrees F.

acknowledgements

Several people contributed in different ways to the success of this project. I am grateful for the assistance of W4MJJ, W4RJV, WA4HSY, and WD4LWL.

conclusion

I would welcome any comments or suggestions. Please include an SASE if a reply is desired. Printed circuit boards may be available. Contact the author for details.

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ham radio

vacuum tube substitution

Using one tube in place of another saves time, trouble

Radio Amateurs always keep a supply of spare parts and vacuum tubes on hand, but when a vacuum tube fails and an exact replacement is unavailable, the Amateur has two choices: either to buy a new tube — which may be expensive or scarce — or find a substitute.

Because a substitute tube may be cheaper than the original, and perhaps easier to obtain, this article discusses tube substitution in general and rectifier tubes in particular.

Unfortunately, tube substitution handbooks aren't always very useful. This would not be so if most of the substitutes were listed, but this is seldom the case. For some tubes, no substitutes are listed; for others, suggested substitutes denoted by suffixes such as *G, GT,* and *GTA* may differ only slightly from the original and be as costly or as difficult to obtain. Sometimes substitutes may be as hard to find as the originals themselves. (Receiving tube manuals may be more helpful in that they list many tubes which have identical base diagrams and are thus interchangeable if they have similar or superior characteristics.)

Tubes with the same kind of base but with different terminal diagrams require only minor changes in socket wiring for substitution. Tubes with the same kind of base and known terminal diagrams, but different filament or heater voltage, may also be used; a higher voltage requires a different filament transformer or secondary winding, and a lower voltage requires a dropping resistor in series with either one of the two socket terminals. The resistance of the dropping resistor is determined by Ohm's law: the drop in voltage is divided by the current in amperes. The power rating must exceed the product of these factors, or E·I watts. Current differences are seldom important.

If a substitute tube has a different base than the *original,* it requires a very different socket unless a suitable adapter is available. It is both difficult and impractical to change bases on old tubes; the absence of separate bases permits only socket rewiring for substitution.

tube-base history

Early vacuum tubes were often baseless or had only bases for filaments. Candelabra bases, Edison medium screw bases, and Ediswan bases were common. From 1915 to 1927 various 4-pin bases and sockets were tried, with varying success. In 1923 the famous UV199 and UV201A triodes were introduced. Their UV bases had solder-filled brass pins which made butt-end contacts with flat socket contacts. The solder tips often corroded badly, so some tips were gold-plated and later, in 1925, UV bases were superseded by UX bases with long pins for side-wiping contacts with ordinary metals.

In 1927 the 27-triode detector-amplifier was introduced to avoid hum in new AC receivers. It had a unipotential or heater-type cathode and required a 5-pin base. The then-new 24A screen-grid tube or tetrode, also of heater-type, needed a 5-pin base and a top cap for connection to the control grid. The advent of pentodes, beam power tubes, and multifunction tubes led to 6- and 7-pin bases and often top caps. The use of 4- to 7-pin bases practically ended in 1935 when octal-base tubes were introduced.

Most tubes required no more than seven active

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pins, but the 6L7 pentagrid mixer required all eight pins plus a top cap, for a total of nine connections. Miniature baseless tubes, which began to replace octal tubes in about 1950, are usually 7-pin or 9-pin types. Some multi-unit tubes have ten or twelve pins.

The first vacuum tube rectifiers on the market were the half-wave UV216 in 1921, (replaced by the UX216B in 1925), and the full-wave UX213 in 1925, which was superseded by the UX280, (now simply 80), in 1927. These tubes, sold by RCA, were developed by GE and Westinghouse, respectively. Unlike other types, rectifier tubes largely survived the trend to more than 4-pin bases from 1927 to 1935. Octal-base rectifier tubes have endured over the years, as demonstrated by the presence of approximately fifty rectifier tubes in the octal-base list (**table 1**). Notable exceptions to the 4-pin and 8-pin rectifier preference are the 5-pin 84/6Z4, which had a heater-type cathode, and the 6-pin 25Z5, which had a heater, two cathodes and two plates, and provided doubler operation. Half-wave rectifiers need only three connections, of which two are for filament and one is for plate — usually a top cap, so only two base pins are wired. Full-wave filament-type rectifiers need four pins — two for filament and two for plates.

rectifier type substitution

There are several good reasons why rectifier tubes are usually the first tubes to fail or weaken. They sup-

table 1. Rectifier tube base diagrams. Commas signify common connections. Column *K* dashes signify filament/heater-only type tubes, with no cathode employed; column *K* numbers indicate cathode element pin connection(s). (M) signifies mercury vapor; TC, top cap plate connections (half-wave).

pins	diagram	tube type	F or H	K	P
4	4AB	2X2A, 2Y2	1 + 4	–	TC
	4AD	83V	1 + 4	4	2 + 3
	4AT	872A/872 (M)	2 + 4	–	TC
	4B	2Z2/G84, 866 Jr. (M)	1 + 4	–	2
	4C	5X3, 5Z3, 80, 82 (M), 83 (M)	1 + 4	–	2 + 3
	4G	6Z3	1 + 4	3	2
	4P	3B28, 816, 836, 866A, AX, B(M)	1 + 4	–	TC
5	5D	84/6Z4	1 + 5	4	2 + 3
6	6E	25Z5	1 + 6	3 + 4	2 + 5
7 (none)					
8	3C	1G3GTA/1B3GT, 1K3A/1J3	2 + 7	–	TC
	5DA	5AR4/GZ34, GZ32-34-37, R52, U54, U70	2 + 5	5	3 + 4
	5DE	3DG4	1 + 3	–	5 + 7
	5L	5AT4, 5V4GA, 5Z4, GZ32	2 + 8	8	4 + 6
	5Q	5X4G, 5Y4G-GA-GT	7 + 8	–	3 + 5
	5T	5AS4A, 5AW4GT, 5AZ4, 5R4GB, 5R4GY, 5T4, 5U4GB, 5V3A/5AU4, 5W4GT, 5Y3GT, 5Z10, GZ30-31, RJ2, U50-51-52, WTT102	2 + 8	–	4 + 6
	6AD	35Z5GT, 45Z5GT	Midtap 3, 2 + 7	8	5
	6S	6AX5GT, 6X5GT, 6ZY5G, EZ35	2 + 7	8	3 + 5
	8EZ	3A3C, 3CZ3A	2 + 7	7	TC
	8KS	5DJ4	1, 7 + 2, 8	–	3, 4 + 5, 6
	8MK	3CU3A	2 + 7	–	TC
	8MU	2CN3A, 3CN3B	2 + 7	7	TC
	8MX	3DB3/3CY3, 3DJ3	2 + 7	1, 7	TC
	8MY	3DA3/3DH3	3 + 5	8	TC
	8MZ	3DC3	2	1, 3, 5, 7	TC
7 min.	5BQ	35W4	Midtap 6, 3 + 4	7	5
	5BS	6BX4, 6X4, 12X4, EZ90, EZ900 HZ90, U78, U707, VZM70, 6202	3 + 4	7	1 + 6
9 min.	9BS	12DF5	4 + 5	3 + 8	1 + 6
	9CB	6AF3, 6AL3	4 + 5	TC	9
	9DJ	6BW4	4 + 5	9	1 + 7
	9DT	3A2A	2, 5, 8	1, 4, 6, 9	TC
	9M	6CA4, EZ4, U709, UU12	4 + 5	3	1 + 7
	9NT	5BC3	1 + 3	–	5 + 9
	9QT	5BC3A	1 + 2, 3	–	5, 6 + 8, 9
	9U	1V2	4 + 5	–	1, 9
	9Y	1X2A, B, C	2, 5, 8	–	TC

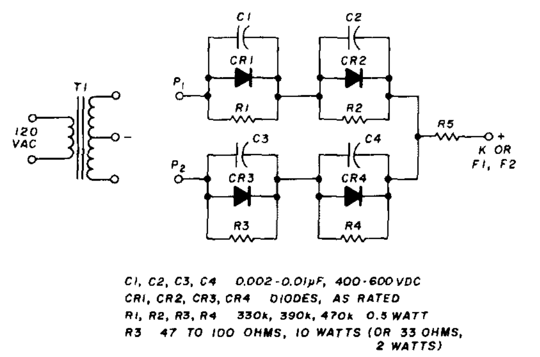


fig. 1. Solid-state rectifier top connections. See table 1 for base connections.

ply DC plate power for other tubes in radio receivers and for much electronic equipment. They also frequently sustain severe transients and momentary overloads. While many suitable substitutes are available, many of these require changes in socket wiring, as evident in **table 1**, which lists about fifty rectifier tubes among fifteen octal-base diagrams. **Table 1** includes both foreign and domestic half-wave and full-wave rectifier tubes. It will assist in substituting one tube for another as well as substituting solid-state equivalents for rectifier tubes.

solid-state rectifier substitution

It is unnecessary to substitute tubes or to change socket wiring or sockets if solid-state parts are mounted in an old vacuum-tube base or a similar plug foundation. (The bases of defunct tubes are easily removed by hack-sawing them about midway or by crushing the glass in a vise. A cloth wrapper will help prevent the hazard of flying glass particles.)

Each leg of a full-wave rectifier circuit must withstand a peak inverse voltage of 1.4 times the RMS voltage to center-tap of the transformer secondary. This voltage usually exceeds the PIV rating of a single diode, so two or more diodes must be used in series. Each diode should be shunted by a small capacitor in order to minimize voltage spikes, and by a resistor to equalize PIV drops. Capacitors should be rated at 400 to 600 volts and 0.002 to 0.01 μF capacitance. Resistors should be rated at one-half watt minimum and 500 to 1000 ohms per volt of PIV. Values of 330k, 390k, and 470k, are suitable at 200 to 400 volts per diode.

These parts should be chosen carefully and connected properly, according to the diagram of **fig. 1** and the pin numbers of **table 1**. The two anode or plate leads (P1 and P2) and one cathode or filament lead (K or F) must be connected to three specific pins in the desired tube base. Solder in the base pins must be melted so as to pass the leads, which may be No. 18 solid copper wire; the pins must then be resoldered.

It is seldom necessary to use diodes in parallel because their current ratings usually exceed those of rectifier tubes. If two or more diodes must carry the load current, each diode should be in series with an equalizing resistor of about 33 to 47 ohms and 2 watts minimum rating. Without series resistors, one of several nonlinear devices in parallel may take most of the current and burn out.

A limiting resistor of at least 10 watts and 47 to 100 ohms rating may be connected in series with a rectifier output or a transformer center-tap connection. The resistor may burn out under overload conditions, but it will protect the diodes and cost less to replace. A well-chosen fuse may be used instead of, or in addition to, the resistor.

rectifier design data

The following facts deserve consideration in rectifier design:

1. Half-secondaries of plate transformers usually have about 50- to 200-ohms resistance, and power-line inputs to transformerless rectifiers have negligible resistance.

2. Choke-input and resistive input filters usually have more than 100-ohms input resistance, and capacitor-input filters have negligible input resistance until the capacitors are charged.

3. Vacuum-tube diodes usually withstand high peak currents caused by short circuits and starting transients because their plate resistance exceeds 500 ohms and their plate current is limited by filament emission and space charge. Solid-state diodes, on the other hand, may fail because they usually have less than 100 ohms resistance and no current-limiting saturation phenomena.

4. Peak current in all rectifiers may be limited to safe values by adding series resistance of high enough value to make the total resistance adequate but low enough value to avoid an excessive voltage drop under normal load.

5. In order to avoid abnormally-high peak currents and peak inverse voltages, as well as poor ripple reduction, in choke-capacitor filters, the product of the choke inductance in henrys and the capacitor capacitance in microfarads must be greater than the resonant value of 7.08 for a half-wave filter (60 Hz ripple) and 1.77 for a full-wave filter (120 Hz ripple).

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

Today's popular all-solid-state transmitters's would make yesterday's old timers' eyes pop out. What fun to work a modern rig with all the bells and whistles!

Owners of today's transmitters, however, find that many of them require matching to an antenna with an SWR less than 2:1. This is no problem on the higher bands, since most antennas meet this requirement, but 80 meter operators discover that they cannot cover the whole band with a single simple antenna and are therefore restricted to operating within a small portion of the band.

Some heroic attempts have been made to solve this problem[1,2,3] *but* the "broadband" 80-meter antenna solution still seems in doubt — or has it already been solved?

let's not reinvent the wheel

The basic 80 meter, center-fed dipole, mounted in close proximity to ground, has a feedpoint resistance in the neighborhood of 30 ohms. I say "in the neighborhood" advisedly, as the measured resistance varies with height above ground, the conductivity of the soil in the vicinity of the antenna and the degree of coupling between the antenna, the outer shield of the feedline and any other conductors (such as house wiring, telephone lines, etc.) in the vicinity of the antenna.

As a result of these variables, SWR measurements made on one 80-meter dipole may vary considerably from those made on an identical antenna at a different location.

Because of the antenna environment, some "lucky" 80-meter operators find their dipole has an extremely broad frequency response and they can operate their solid-state transmitter over nearly the whole band! But *their buddy across town* with the same antenna is limited in operation to a small segment of the band, since the SWR quickly departs from low values when he operates his antenna away from the design frequency.

Improving ground conductivity is a difficult task.[2] An extensive ground screen is called for in the case of vertical polarization and it is not known if such a ground installation would be cost-effective with horizontal polarization. I would doubt it, myself. [*It would, at the minimum, better define the "array" (antenna plus image) elevation pattern.* — **Editor**]

decoupling the antenna from the environment

Meaningful SWR measurements are difficult to achieve when the transmitting antenna is coupled to nearby conducting objects. In my case, the SWR measurements on an 80-meter dipole changed radically when I turned on the ceiling light fixture in the living room. The dipole *was* parallel to the wires leading from the utility box at one end of the house to the light fixture. Placing a 0.01, 1.6-kV ceramic capacitor across the connections of the fixture seemed to detune the house wiring sufficiently so that meaningful SWR numbers could be obtained.

The first step was to determine if the dipole was coupled to the outside of the outer shield of the coaxial feedline. This was tested by adding an extra length of line at the station end and noting if the SWR reading changed from the original measurements at different frequencies.

Since I knew there was unwanted coupling, I was not surprised when I was able to plot a new SWR curve that had only a vague resemblance to the old one after *my line-splicing* experiment. It looked as if I could move

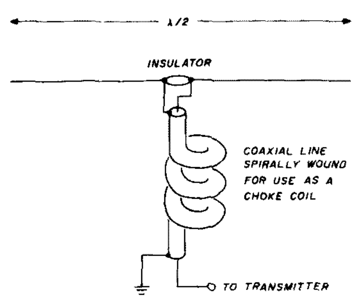


fig. 1. Transmission line wound into simple RF choke at antenna feedpoint helps to decouple outside of shield from antenna currents. The line should be dropped (down) directly under the antennas. Suggested coil diameter: one foot for RG-8A/U, 3 inches for RG-58/U. Eight turns are suggested for use on 80 meters.

the resonant frequency at will within the band by merely changing the length and position of the coaxial line with respect to the dipole!

The second step was to bring the coaxial line directly down to the surface of the ground under the center of the dipole and run it along the ground to the station. Before, it had looped through the air at about a 45 degree angle to the dipole.

Repeating the SWR measurements showed that by varying line length the SWR changed but not nearly as much as in the previous situation. It looked as if I were proceeding in the right direction.

The next step was to try to decouple the outer shield of the coax a bit more. I wound the line up into a choke coil just under the center of the antenna (**fig. 1**). This helped, but it seemed as if more isolation were required. Luckily, I had a 2.5 MHz to 15 MHz air-wound balun available (**fig. 2**). I placed this device at the dipole feedpoint and was successful in decoupling the coaxial shield from unwanted antenna-induced currents.

a balun transformer for 80 and 160 meters

The balun design shown in **fig. 2** is useful for both 80 and 160 meters. Since it has an air core, it will not saturate at a high power level as the ferrite core design might do; and because the windings have more turns (inductance) than the more common design, this balun performs better at the lower frequencies.

The balun is wound on a plastic (PVC) form 3-1/2 inches (9.0 cm) in diameter. The design consists of 10 trifilar turns on No. 14 Formvar™ (or enamel) insulated wire. The ends are held in place by 4-40 hardware. The windings are interconnected by short lengths of wire run between the appropriate terminals. The common connection of two of the windings is used as the ground point at one end of the balun and is attached to the coaxial shield. When completed, a plastic bottle is cut to fit over the balun as a rain shield. The balun is attached directly to the center insulator of the dipole and the coaxial line dropped down directly beneath it.

Take care that the top end of the coaxial line is sealed from moisture. Water can seep into the line by capillary action of the shield but a good coat of sealant (RTV, for example) will waterproof the end of the line. (Make sure your sealant does not contain acetic acid, or it will corrode the copper wires of the coax. Read the label before you buy.)

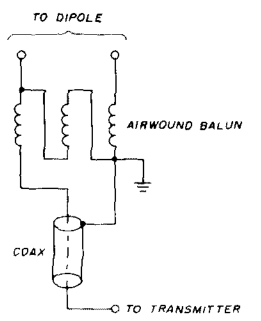


fig. 2. Airwound, trifilar balun covering the 1.8 MHz to 12 MHz range. Ten closely-spaced trifilar turns of No. 14 enamel wire are wound a 3-1/2 inch (9 cm) diameter form. See text for details.

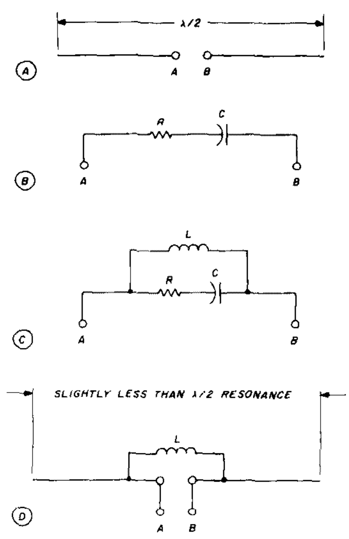


fig. 3. (*A*) Resonant dipole presents purely resistive feedpoint. (*B*) Dipole slightly shorter than resonance presents capacitive reactance in series with feedpoint resistance. (*C*) Inductor placed across B network provides impedance transformation to 50 ohms by simple network made up of R, L, and C. The L/C ratio determines transformation ratio when LC is resonant at operating frequency. (See *Beam Antenna Handbook*, Radio Publications, Box 149, Wilton, Connecticut 06897.)

matching the antenna to the line

I was now in a position to make a meaningful SWR measurement of the antenna. As expected, a minimum of 1.56:1 occurred at resonance corresponding to an antenna impedance of 32 ohms. The easiest way to achieve a better match between antenna and line is to make the antenna form a portion of a network whose input impedance over a small range is 50 ohms (**fig. 3**). If the antenna is cut slightly shorter than its resonant length, its terminal impedance will have a capacitive reactance term. A compensating inductor placed across the antenna terminals would then provide a tuned network whose total impedance can be made to match the coaxial line impedance. For an antenna (radiation) resistance of 32 ohms,

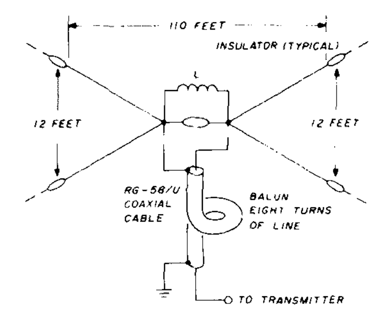


fig. 4. Broadband fan dipole for 80 meters. Wires may be either in horizontal or vertical plane. See *fig. 3* for data on coil L.

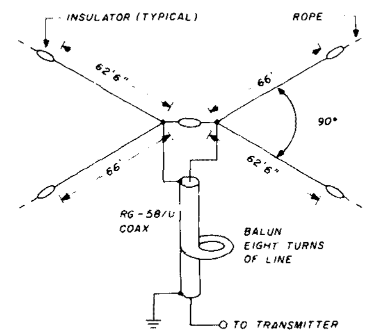


fig. 5. Parallel-fed dipoles placed at right angles to each other provide wideband response. Dipoles are cut for opposite ends of the 80-meter band. See *fig. 6* for coaxial balun data. (Note: Dipoles are in horizontal plane.)

the shunt coil (L) should have a reactance of about 64 ohms. For the 80-meter band, this turns out to be about 2.7 μH. A coil of 8-1/2 turns of No. 12 wire, 2 inches in diameter and 2 inches long will do the job. The antenna is trimmed six inches at a time until unity match is achieved at the desired operating frequency of the antenna.

antenna bandwidth

After this task is completed, what will the operating bandwidth of the antenna be between the 2:1 SWR points? Alas, the correctly-designed dipole still will not cover the whole 80-meter band, falling within the desired SWR ratio limits over only about 220 kHz. But this exercise provides a clue that may help solve the bandwidth problem.

broadband dipoles

The previous discussion illustrates a method of feeding and matching a centerfed 80-meter antenna system. It applies equally well to a broadband antenna as to a conventional antenna.

The fan dipole shown in **fig. 4** provides good bandwidth on 80 meters when properly matched to the transmission line. It is only 110 feet (33.53m) long at resonance (3.75 MHz). When the length of the arms and the center matching coil are properly adjusted, the *antenna* exhibits an SWR of less than 2:1 over the complete 80-meter band.

A second broadband antenna design similar to the fan dipole is shown in **fig. 5**. This scheme consists of two parallel-fed dipoles placed at right angles to each other. The dipoles are cut for opposite ends of the band. (I haven't tried this idea, but I am waiting for a report from someone who has.[3]). The dipole's SWR follows a W curve with points of minimum SWR occurring *near the band* edges. The dipoles have to be physically separated by at least 60 degrees; otherwise, the broadband response is lessened.

Again, as in the case of the simple dipole, the input impedance of the antenna is quite low and a matching coil has to be placed across the feedpoint to provide a step-up transformation to 50 ohms.

The idea looks like a good one and with sufficient separation between the dipoles, it should work.

the W6TC two-frequency dipole

George, W6TC, wrestled with the problem of using a dipole across the 80-meter band and finally came up with a classic solution so simple that I wish I had thought of it myself!

George's idea is shown in **fig. 6**. This illustrates a dipole cut for 3800 kHz and used for SSB operation. To

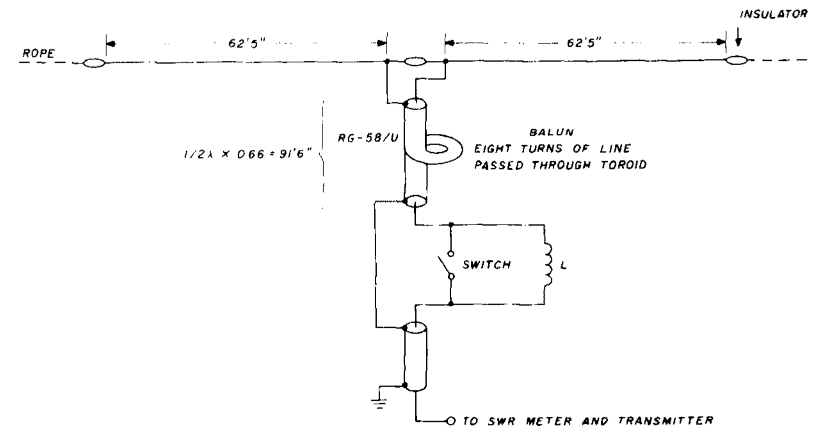


fig. 6. The W6TC two-frequency dipole for 80 meters. Antenna is cut for resonance at high frequency end of band. A single loading coil for operation at the low frequency end of the band is placed an electrical half-wavelength down the transmission line. The line is decoupled from the antenna by a simple balun placed directly at the antenna insulator. Toroid is powdered iron material T-200 (2″ OD) (1.25″ ID), 2-mix (red, μ = 10 (Amidon T-200-2), or equivalent.

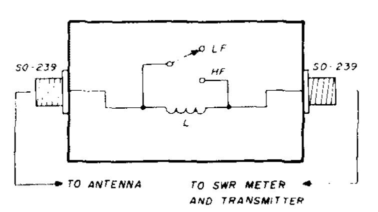


fig. 7. Remote loading inductor is placed in shielded box at operating position. Coil is approximately 11 turns No. 12 wire, 2″ diameter and 2-1/4″ long. (About 3.75 μH). Switch is surplus ceramic unit.

operate at 3500 kHz, George first placed a single loading coil in series with one leg of the dipole. The unbalance caused by not splitting the coil and putting half of it in each leg was unnoticeable.

Now, the problem is that the dipole has to be raised and lowered to insert or remove the loading coil for a QSY from one end of the band to the other. George solved this problem by moving the coil a half-wavelength down the coaxial feedline, placing the loading coil right at the operation position.

Physically, the coil is placed in a box as shown in **fig. 7**. A ceramic switch is connected across the loading coil and instant QSY between the ends of the band is possible — right from the operating position. The two SWR curves for the antenna are shown in **fig. 8**.

another easy way

In microwave antennas, a device known as a "double-stub tuner" can be used to reduce the SWR on a feed system. Simpler versions of this tuner have become known as "line flatteners" and one that will work on 80 meters is shown in **fig. 9**.

Using the line flattener, an 80-meter dipole cut for one portion of the band can be made to work anywhere in the band with near-unity SWR presented to the transmitter. Under the worst of circumstances, the SWR on the line could go as high as 5:1 if a dipole cut for one end of the band were used at the opposite end of the band. Even so, the line flattener can readily take care of the problem. In most instances, the dipole is cut for some frequency within the band and maximum SWR excursions run closer to 3:1 at the band edges. No matter. The line flattener does the job.

The device is easy to adjust. The capacitors and inductors are adjusted until the SWR at the transmitter is unity. If you run out of capacitance range in one unit, a 1250 working volt mica capacitor placed in parallel with the fully meshed capacitor will help.

any ideas from the field?

Well, I've heard of the scheme of making the 80-meter dipole out of steel wire to introduce a little loss and thereby lower the circuit Q and improve the bandwidth. And it might work, but I don't have feedback from anybody who has tried the idea.

But I would like to hear from those who have any original thoughts on an 80-meter broadband antenna system that will show a low value of SWR to the new solid-state rigs.

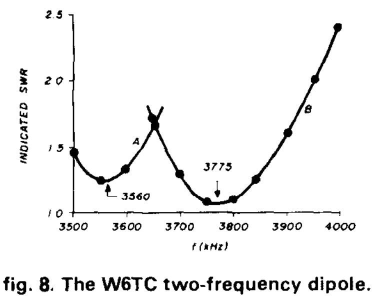


fig. 8. The W6TC two-frequency dipole. Curve A shows resonance at 3560 kHz with a remote loading inductor in the circuit. Curve B shows resonance at 3775 kHz with inductor shorted out. Resonance curve B can be shifted by changing length of antenna; curve A can be shifted by changing the value of the loading inductor.

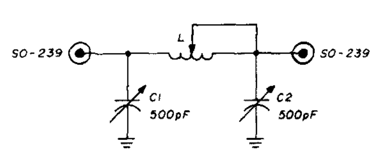


fig. 9. A coaxial "line flattener" for 80 and 160 meters. Capacitors may be shunted with 1250 volt mica units to increase range. (As much as 2000 pF may be required on 160 meters.) Inductor can be tapped or switched. No. 16 bare or tinned copper wire is recommended as a starter (12 turns per inch). Surplus rotary inductor would be useful.

computermania!

In my April, 1983, column I discussed short, loaded dipole antennas and showed a simple computer program that would aid in the fast design of a short dipole for any Amateur band. Missing from the program was the section required to design the impedance matching coil. The program was modified by Dick, W6EDE, for a TRS-80 (II), and has been further refined by John McComic, K4KAJ (**fig. 5**). John's program covers the complete antenna design, plus matching coil, for any frequency. The result is a half-length dipole, perfect for cramped locations.

A copy of the missing section of the program is available directly from *ham radio*. Be sure to enclose a stamped (20¢) business-size envelope (9-1/2 × 4-1/8) with your request.

(My thanks to the following who also provided interesting and useful antenna programs: Lloyd Phillips, WB6WCA; Warner Thompson, N7WT; and I. L. McNally, K6WX.)

references

1. William I. Orr, W6SAI, "Ham Radio Techniques," *ham radio*, August, 1982, page 42. (See Snyder 80-meter antenna product review in this issue. **Editor**)
2. Jerry Hall, K1TD, "The Search for a Simple Broadband 80-Meter Dipole," *QST*, April, 1983, page 22.
3. Mason A. Logan, K4MT, "Stagger-Tuned Dipoles Increase Bandwidth," *ham radio*, May, 1983, page 22.

ham radio

short circuits

vertical phased arrays

The following corrections should be made to part 3 of K2BT's article, "Vertical Phased Arrays" (July, 1983):

Eq. 3 (p. 30) should read:

$$Z_{12} = \pm \sqrt{Z_{22}(Z_{11} - Z_1)}$$

The polar notation in the upper left-hand column of page 32 should include the angle symbol and read:

$$20.4 \angle 132.2^\circ$$

The following lines identifying the driving point impedances in table 3 (page 33) should be corrected as indicated.

3-element in-line array, λ/4 spacing:

Z_1	−6.6 − j21	should read 15.4 − j17
Z_2	51.4 + j0	should read 36.2 + j0
Z_3	79.4 − j39	should read 75.4 + j43

triangular array, 0.289λ spacing:

Z_1	28.4 − j10	should read 20.4 − j10

panoramic adapter

The following corrections should be made to "Design Notes on a Panoramic Adaptor/Spectrum Analyzer" by Rick Ferranti, WA6NCX, (February, 1983):

In fig. 4 (page 30), the connection shown between the transistor's collector and the 3.3k resistor is erroneous. The collector lead is still attached to T1.

In reference to the 31 MHz bandpass filter in fig. 8 (page 31) the notation "space coils 1/2 inch apart . . ." refers to the distance between T_1 and T_2 (centers); the 2-turn windings on the coil should be over the 12-turn windings on each form at the cold end.

A small variable capacitor across the 4.7k resistor was deleted from fig. 9 (page 32), but not from the text. Because the capacitor pulled the oscillator only slightly in frequency, it is not necessary in this circuit.

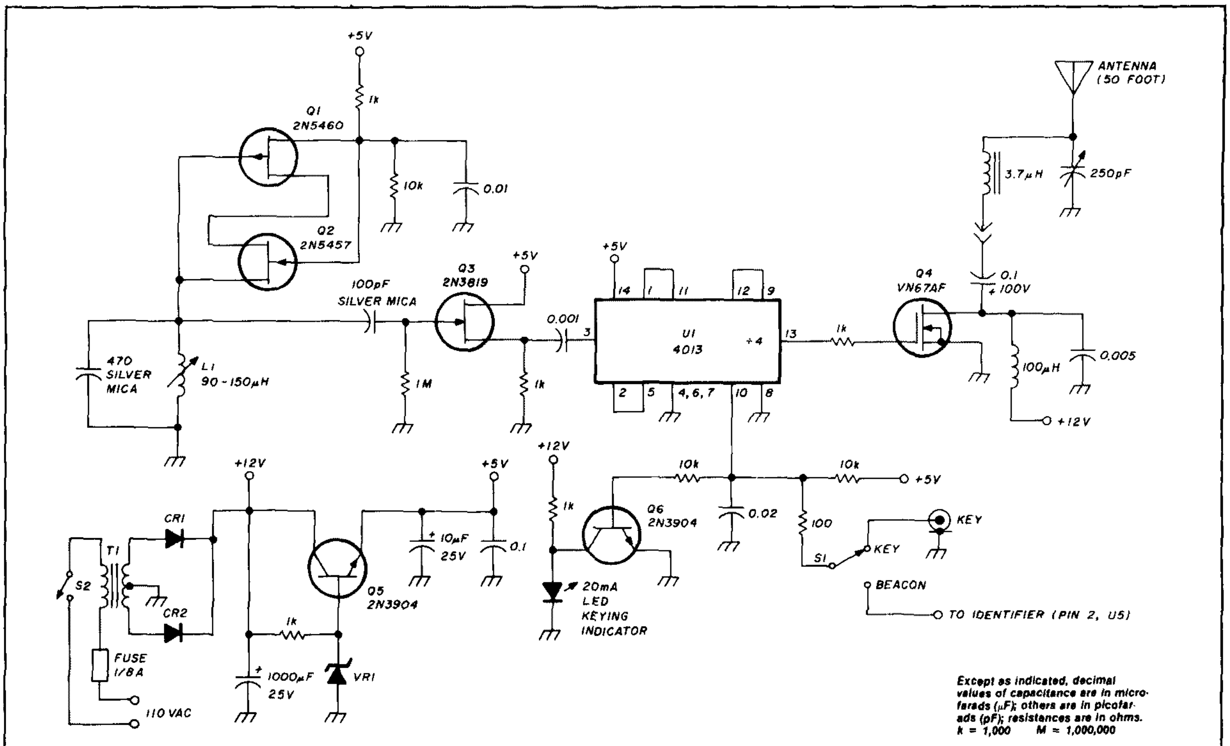


fig. 1. Schematic diagram of the LF transmitter. The VMOS final amplifier consumes 86 percent of the total current, running less than 1 watt input. (CR1, CR2 are 1N4002 diodes, VR1 is a 1N52 Zener diode (5.2V), and T1 is 110 Vac primary, 24 V ct secondary, 300 mA transformer.

VMOS on 1750 meters

A CW or beacon transmitter for 160 kHz

Many interesting experiments and enjoyable QSOs can be your reward for operating on the 1750-meter band. No license is required to use the slice of radio spectrum between 160 and 190 kHz, and it requires only simple equipment to provide the legal maximum of 1-watt input to the transmitter. Enthusiasts have made contacts over distances of 700 miles in spite of the antenna-size limit of 50 feet.

I've used my VMOS transmitter to make a contact from East Haven, Connecticut to Owings, Maryland — a span of 250 miles. It can be done consistently as long as very noisy conditions do not exist. Many schedules have successfully been kept with stations in New Jersey, New York, Maryland, and Massachusetts. My beacon signal has been heard as far north as New Hampshire and as far south as Maryland. Receiving equipment for this band is not at all difficult to build.[1]

D-layer reflection and groundwave propagation modes dominate the band and best results are obtained by installing a good ground system and by keeping antenna coupling-losses low. Radiation resistance for the 50-foot antenna is on the order of 0.02 ohms and the radiation efficiency is therefore very low. Nevertheless, a few milliwatts of effective radiated output does give you usable communications.

By S. J. DeFrancesco, K1RGO, 17 Jeffrey Road, East Haven, Connecticut 06512

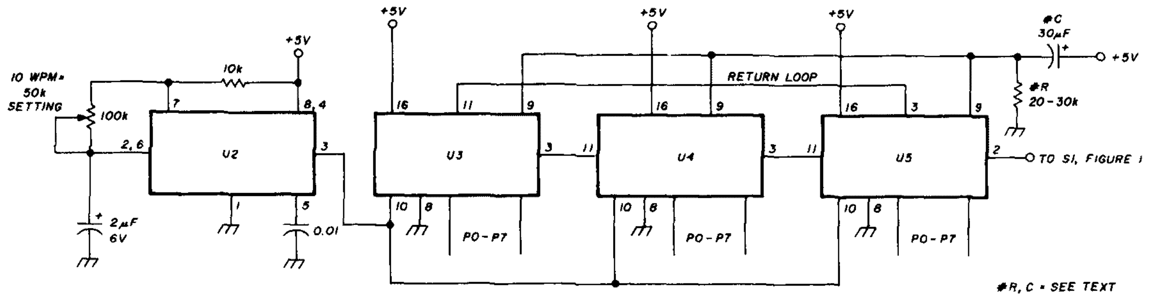


fig. 2. Three universal shift registers and a clock-pulse generator make up a programmable, 24-bit, closed-loop identifier. U2 is an NE555, U3, U4, and U5 are 4021 eight-bit shift registers.

the transmitter

The transmitter circuit diagram shown in **fig. 1** uses a 600-800 kHz negative-resistance oscillator consisting of Q1 and Q2. L1 is a variable inductor of 90-150 μH. The 470 pF capacitor can be made variable, and an old BC loop stick can be used for L1 instead. Stability is excellent — only a few hertz of drift were observed one hour after the unit was turned on. Isolation is provided by Q3, a source-follower circuit which drives U1. U1 is a CMOS 4013 type-D dual flip-flop used to divide the 600-800 kHz from the oscillator down to 150-200 kHz. The output of U1 drives the VMOS amplifier. The output of Q4 is coupled to the antenna tuning network through a 0.1 uF capacitor. The antenna tuning network consists of a 250 pF variable capacitor and a 3.7 mH coil wound on a ferrite rod. This high *Q* L/C arrangement resonates the 50-foot vertical antenna to 160-190 kHz.

Transmitter keying is accomplished by gating U1 on and off at pin 10. Q6 is an LED driver connected to indicate keying. S1 is set to key position for manual CW operation, or BEACON position for continuous-message operation.

identifier

Three 4021 CMOS universal eight-bit shift registers are used in the identifier for minimum power consumption. In **fig. 2**, U3, U4, and U5 clock inputs are connected in parallel. U2 is a square-wave clock-pulse generator which drives the other ICs. The 100k potentiometer can be adjusted to obtain the desired code speed. The P0-P7 parallel inputs are programmed to provide identification. R* and C* are somewhat critical and should be selected for the code speed used. R*, shown in **fig. 2**, is 30k for 15 wpm or less, or 20k for 20-30 wpm. C* is 30 μF for this same range of code speeds.

When power is first turned on, P0-P7 parallel inputs are loaded in serial with pin 9 on U3, U4, and U5 in a logic-high state. As C* equalizes, and pin 9 goes low, the serial-loop mode is activated and the program is stored. It then circulates in a continuous loop. The output (pin 2 of the U5) pulse train is fed to

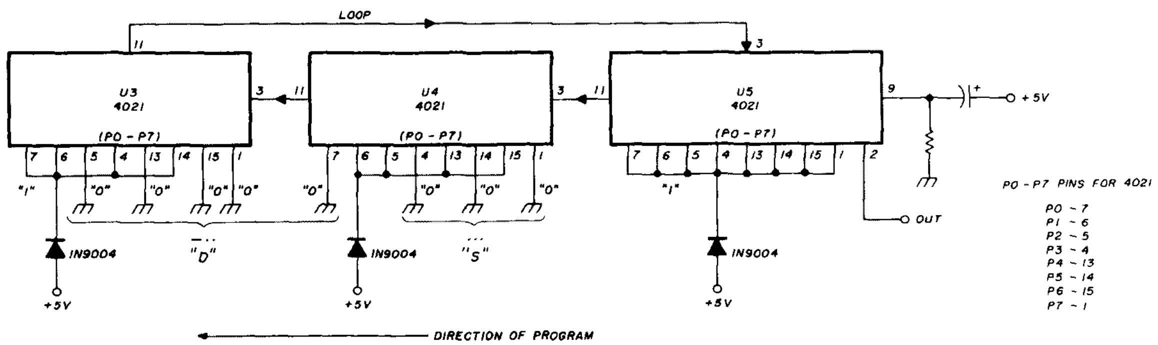


fig. 3. The identifier circuit is shown programmed for the letters SD. Calculate the number of bits you need for your identifier, and start at U4. Allow at least three space bits to separate the character start.

U1 in the transmitter, providing CW transmission in the beacon mode. Code speed is varied by changing the output frequency of U2.

programming

A dash is equal to three bits (zero) and a dot is equal to one bit (zero). Spacing between dots and dashes to form a character is one bit (zero). Spacing between characters is two bits (1). The identification I use is "SD," consisting of fourteen bits with ten bits to spare. The program is shown in **fig. 3**, using the P0-P7 parallel inputs. A 0 is an on bit and a 1 is an off, or space, bit. Once the program is wired in, no other preprogramming is needed. If your program appears to drop a bit or two, shut off the power for a few seconds and then switch it back on again; this will reset the program.

operation

I monitored the transmitter on an RBA-5 low-frequency receiver, and adjusted the VFO to the desired frequency (185 kHz). When I keyed the transmitter, I heard very little key clicking, and a nice, chirp-free signal was generated. I tuned the output capacitor for maximum indication on my field strength meter. An NE-2 neon bulb glowed brightly when touched to the antenna terminal. I then tuned the 250-pF loading capacitor and found that no shift in frequency was being caused by the antenna tuning. I turned the switch to the BEACON mode to evaluate the IDer, and it worked fine.

This system has run twenty-four hours at a time and has been very reliable. Total current drain is 100 mA, of which 86 mA is used by the VMOS power amplifier. Input to the final is around 1 watt, the legal unlicensed limit.

construction notes

The complete unit can be made very small, including the power supply — it is possible to fit everything except the antenna loading circuit into a 2 × 4 × 2-inch (50 × 103 × 50 mm) Minibox.

The 3.7 mH antenna coil can be air core or low-loss ferrite core. I close-wound 138 turns of No. 26 wire on an old 5/8-inch diameter broadcast-band ferrite core. A 250 or 365 pF variable capacitor can be used with this inductor to resonate the 50-foot vertical antenna. Experimentation with various cores and coils to improve efficiency in the antenna circuit can be evaluated by field-strength readings.

ham radio

references

1. DeFrancesco, S.J., K1RGO, "A Fixed-Tuned LF Converter," *ham radio*, January, 1983, page 19.

technical forum

Have a technical question? Ask your fellow Amateurs through Technical Forum. You'll reach knowledgeable readers ready to help you find the answer you're looking for. *Have an answer?* Send it in. Every month we'll award a specially selected book from Ham Radio's Bookstore to the writer of the best response to a question previously posed in this column. Address your questions and answers (typewritten, if possible) to Technical Forum, *ham radio magazine*, Greenville, New Hampshire 03048. *The opinions expressed in this column are those of the individual authors and do not necessarily represent the views of the publisher or staff of* ***ham radio magazine***.

pacemakers and RFI: safety first

We read Edwin M. Hollis's letter (Technical Forum, June, 1983) concerning the use of transmitting equipment by a ham with an implantable *cardiac pacemaker*.

There is a general misconception that everyone with a pacemaker will "expire" when exposed to an RF field or a microwave oven. The twenty-five or more manufacturers of pacemakers worldwide are concerned about the response of their products to electromagnetic interference and do extensive testing on the products before release. These tests are done by the manufacturer and by independent testing organizations. Virtually all manufacturers include filters and/or shielding of the electronic circuit.

Pacemakers are prescribed for patients who have some abnormality in the normal electrical activity of the heart. Without this normal electrical activity, the heart will slow down or skip beats entirely. The pacemaker monitors this activity and generally puts out a stimulating impulse only when needed. The stimulating impulse "shocks" the heart muscle, causing the muscle to contract and the heart to beat.

The circuit that monitors the heart activity also looks for interfering signals (EMI, RFI, etc.). If interfering signals are detected, the pacemaker's stimulating pulses will be turned on, instead of shutting off completely. There is a remote possibility of the pacemaker's being *inhibited (turned off)* if the field is intense. (This warning is included in the patient handbooks and physician manuals.) If this were to occur, the patient might experience dizziness or fainting spells. As soon as the interfering signals were removed, the pacemaker would return to normal operation. No permanent damage would be done to the pacemaker's electronic circuit.

In recent literature, there have been no reported deaths due to RFI exposure of patients with implanted pacemakers. There have been reported cases of patients experiencing fainting spells or dizziness with older pacemaker models that are no longer manufactured.

If a patient is concerned about exposure to RF fields, the pacemaker manufacturer should be contacted. The address and phone number are printed on the permanent plastic I.D. card given to all pacemaker patients.

Any Amateur with a pacemaker should follow these general rules:

1. Make sure transmitting equipment is in good working order.

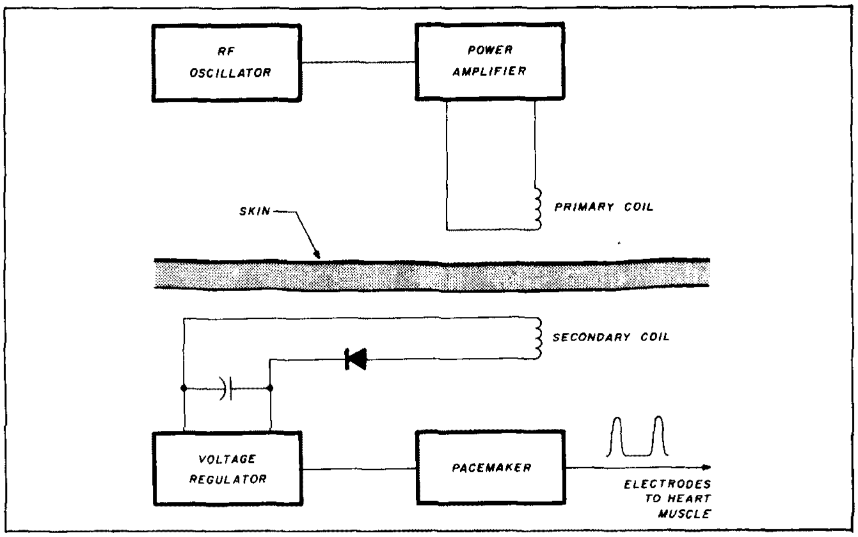
2. Maintain a good ground on all equipment.

3. Avoid working on equipment with cover or shields removed.

4. Do not stand near *antenna* systems radiating power, especially in front of directional arrays.

If you must do any of these things, have someone else nearby to turn off transmitting equipment if any dizziness or fainting spells are experienced. Electrical shock hazard is not increased because of the pacemaker. There is no technical reason why an Amateur Radio operator with an implantable cardiac pacemaker cannot continue to enjoy the hobby by simply exercising a few cautionary measures.

John W. Bixler, K3EAS
Field Program Coordinator
William L. Johnson, K3FOW,
Electronics Manager
Donald A. Henry, W3FE, Engineer
Cook Pacemaker Corporation
Leechburg, Pennsylvania 15656



concern at high levels

I have some information for Ed Hollis, K4CN, on the effects of RFI on pacemakers (Technical Forum, June, 1983).

It depends on the type of pacemaker. Some are RFI powered externally (**fig. 1**); in these, a secondary coil rectifies the induced voltage to power the pacemaker. There are also nuclear-powered long-term pacemakers, employing a *shielded* radioactive source, which are implanted in the patient. The most popular pacemakers at present contain a lithium

battery which can last up to ten years.

Basically, a pacemaker simulates the periodic electrical pulse which stimulates the heart muscle to a state of contraction. Its circuitry can be a monolithic multivibrator or a blocking oscillator of either the asynchronous or the more complex-demand type used in synchronous pacemaker. The latter depends on ventricular contractions' return pulses which automatically adjust the pulse rate with variations in physical activity. The pulse width for pacemakers varies from 0.5 to 2 ms with amplitudes of 5 to 9 volts.

In the normal ham shack you need not worry about any disruption of the pacemaker operation. *But very high levels of RFI can override the pulsed output*, especially affecting the demand type synchronized pacemaker, which is controlled from the heart muscle. Avoid standing under a broadcast station antenna (running kilowatts), stay away from high ERP microwave dishes, and when adjusting your beam antenna's gamma match, *shut your transmitter off*. — **Salvatore J. DeFrancesco, K1RGO**

heartwarmer

Dear HR:

I can think of no adequate way of thanking *ham radio* for printing the item "pacemakers and RFI" for me in your June issue (Technical Forum, page 98). The response to the item has been just overwhelming; the value of information gained from the data and letters cannot be adequately measured. The many letters describing personal experiences and offering technical advice were incredible.

It proved several things: first of all, that a lot of Amateurs subscribe to *ham radio*; and what is more important, they *read* it. If one ever did have any doubts about the type of person the Amateur is, such a response from so many who have taken their time and money to inform another ham (who they never even knew) about a subject which might make the difference between life or death was heartwarming, to say the least.

From the mail I received it is apparent that the pacemaker, along with other forms of medicine and electronics, has come a long way in the past few years. Even though an Amateur himself is not wearing one of the gadgets, it is nice to know that he will not be endangering a neighbor who might come too close to a transmitting antenna, causing an unexpected problem . . . please thank all the readers who spent their time and money to respond to my question.

The tremendous response to the item also proves without a doubt that there has been and just may still be a vital problem in this area. I'm sorry if I gave the impression that the experience I had in the hospital was recent. It was, in fact, quite some time ago and I have been convinced that many changes have been made since then, as well one could expect. In spite of that I found many not convinced as to the reliability of pacemakers in the presence of RFI; I was also unable to find a recent schematic, which is understandable.

Edwin M. Hollis, K4CN

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ham radio has prepared a list of recent articles on possible RFI interference with pacemaker operation. We would be happy to make that list available to concerned readers or their physicians. Address requests to: PACEMAKERS, *ham radio magazine*, Greenville, New Hampshire 03048. (Enclose large SASE.) — **Editor**

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the ham notebook

zero-beat indicator for RTTY

I built the ST-5 demodulator described *in ham radio*[1] but had difficulty finding the zero beat of the incoming signal on my receiver. The ST-5 has a meter for tuning. When the meter reading is identical for MARK and SPACE, the Model 19 machine provides solid print. However, I found it quite difficult to tune in the signal so that the meter would indicate this condition.

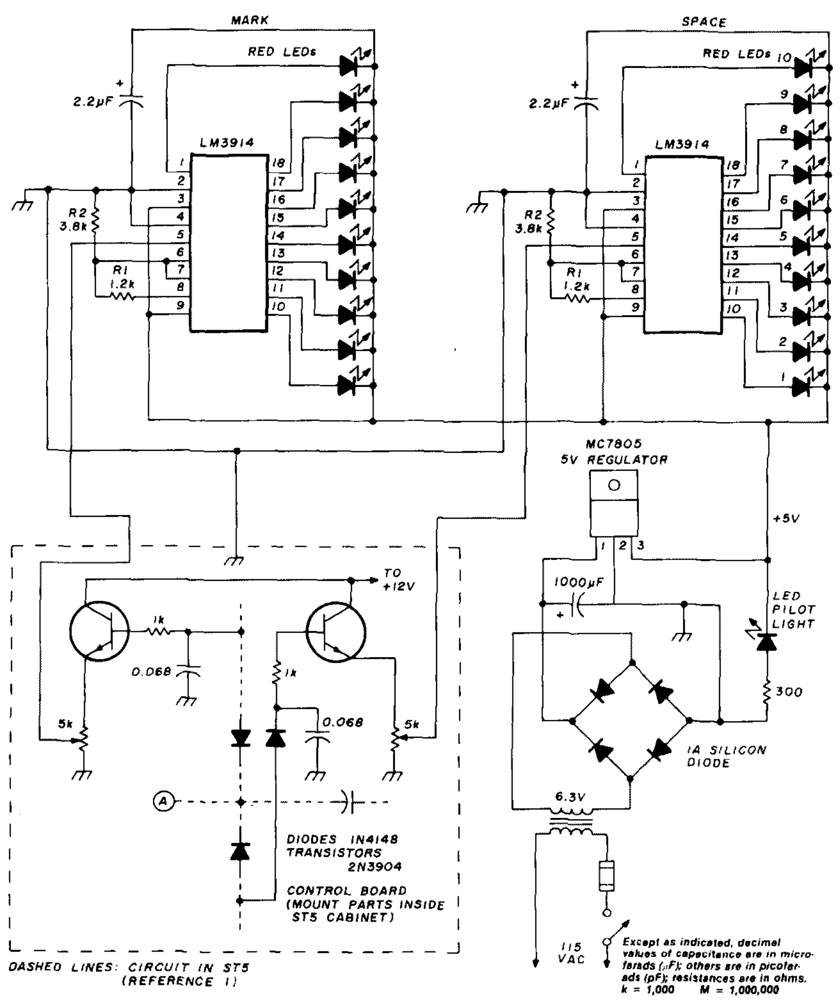


fig. 1. Schematic diagram of the simple zero-beating indicator for RTTY using LEDs and a "bar-generator" IC.

Dave, N2DS, our local solid-state wizard, suggested I make a device using the LM3914 chip* with its ten-LED output. The LM3914 has ten positions for the LEDs, and the resistors for them are built into the chip. The LM3914 is known as a "bar analog generator."

I developed a plan to have two vertical columns of ten LEDs each. One column would be for MARK and one for SPACE. Required were two LM3914s, a 5-volt regulated power supply, twenty red LEDs, and two resistors, R1 and R2 (**fig. 1**).

Resistor R1 is 1.21k. I made this resistor using 1k, 200 ohms, and 10 ohms in series (watch tolerances). Resistor R2 is 3.83k. I used a 3k resistor plus an 800-ohm and a 30-ohm resistor in series. You'll need two R1 and two R2 resistors.

Turning this idea into a construction project resulted in a nice visual device to tune in an RTTY signal. The circuit is more economical and better than an oscilloscope with the high voltage involved and its many controls.

construction

Only three wires are needed to connect the zero-beat indicator to the ST-5. The control circuit (**fig. 1**) can be built on perf or PC board, which is mounted inside the ST-5 cabinet.

Mount the following components onto the control board: two 2N3904s, two 5k trim pots, two 1k ¼-watt resistors, two 1N4148 silicon diodes, and two 0.068-μF disc ceramic caps. Mount the control board in the ST-5 demodulator so that you can adjust

*Radio Shack part #276-1767. LEDs are part #276-1622.

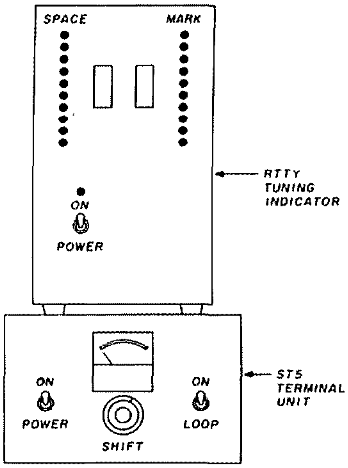


fig. 2. Author's zero-beat indicator mounted on the ST-5 demodulator. Other physical arrangements are possible. Use your ingenuity.

the trim pots easily. In **fig. 2** you'll note that I've mounted the LM3914 chips between the vertical columns of LEDs. No big deal. Perhaps you would like to have the LEDs running horizontally. In any event, if you want to use the vertical display, a better way of mounting the LM3914s is to the left of the LEDs, as shown in **fig. 1**. This makes for easier wiring.

adjustment

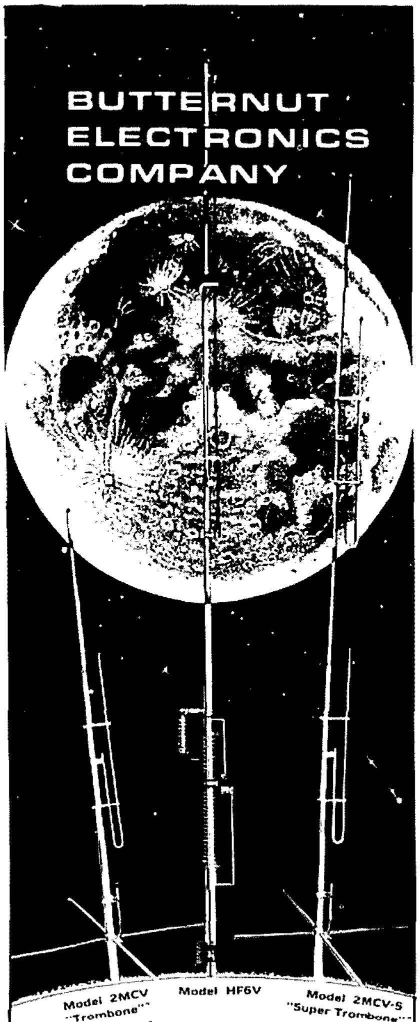
With an input signal, adjust the MARK trim pot so that eight of the ten MARK LEDs light up. Then, with a SPACE signal, adjust the SPACE trim pot so that eight of the SPACE LEDs light. Another method is to tune in W1AW on 3625 kHz. Then make your adjustments, as you'll have a good MARK and SPACE signal. Be sure you get the first 15 minutes of W1AW's transmission; it changes to ASCII during the second 15 minutes.

I'd like to thank Dave Schmarder, N2DS, for his help in completing this project.

reference

1. Irvin M. Hoff, W6FFC, "The Mainline ST-5 RTTY Demodulator," *ham radio*, September, 1970, page 14.

Jim Dates, W2QLI



hf hybrids: a second look

Why not try hybrids in hf applications?

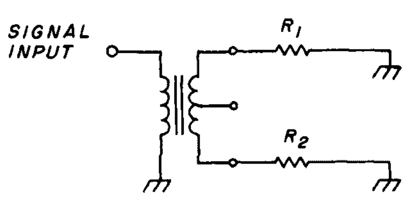


fig. 1A. 180 degree hybrid equivalent circuit.

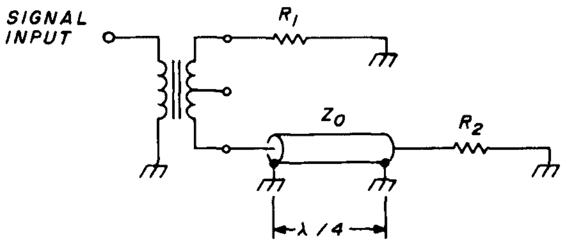


fig. 1B. Quadrature hybrid equivalent circuit.

Hybrids are generally four-port devices used in UHF and microwave designs, where they are easily fabricated as part of stripline or microstrip assemblies. They usually fall into one of two general classes: the 90 degree or quadrature hybrid, and the 180-degree hybrid. Examples of the quadrature type are the side-coupled directional coupler, the branch coupler, and capacitively coupled hybrid.[1] The 180-degree type is probably best exemplified by the ring or "rat race" coupler.

As one might imagine, the quadrature hybrid divides its signal between two outputs which are 90 degrees out of phase with respect to each other. The rat race, on the other hand, provides two outputs which are in phase opposition.

A major difference between the two classes, other than the phase relationship of their output signals, is that in order to obtain a null at the fourth port, the quadrature hybrid requires that the two loads must both match the *impedance level for which the device* was designed. The 180 degree variety, on the other hand, requires only that the two output loads *match each other.*

why?

If we replace our hybrid with yet another and simpler form, a center-tapped transformer, with loads on opposite ends of the secondary coil, (**fig. 1A**), then as long as the two loads are equal, no signal is present at the center tap, which becomes our fourth port. But suppose we require a quadrature phase relationship between the two output signals. We then introduce a quarter wavelength of line between one output and its load. This quarter wavelength of line inverts the impedance of that load, so that unless both loads match the impedance of that section of line, no balance will exist, and a signal will appear at the center tap (**fig. 1B**).

This simple exercise demonstrates why quadrature hybrids must match the loads to the design impedance, while 180 degree hybrids need only to match one load against the other.

The transformer form of hybrid is found to have numerous applications in the high-frequency range. With appreciable power levels, the signal may be introduced into the center tap of the transformer. Dividing into two paths, with opposing magnetic fields, saturation of a ferrite core would be avoided, while the other coil (now the secondary), will show, at its output, any unbalance between the two loads.

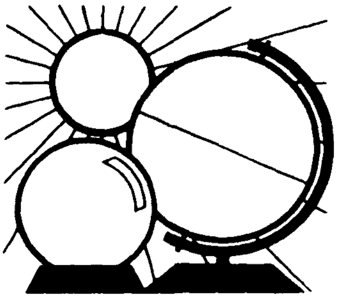
The use of such transformer-type hybrids in the high-frequency range has not been generally exploited, but would seem to offer much for compact power dividers, combiners and phase shifting devices.

reference

1. Henry S. Keen, W2CTK, "High Frequency Hybrids and Couplers," *ham radio*, July, 1970, page 57.

ham radio

By Henry S. Keen, W5TRS, Fox, Arkansas 72051



DX FORECASTER

Garth Stonehocker, KØRYW

trans-equatorial DX

The autumnal equinox — with its promise of winter DX conditions to come — also signals the return of trans-equatorial openings, with their 5000 to 7000 mile (8000-11,200 km) paths.

A cross-section representation of the ionosphere along the 75-degree west meridian is shown in **fig. 1**. Data for this drawing was obtained by ionosonde returns off the bottom half of the ionosphere. Note a concentration of contour lines near the center and right (southern latitude) side. The thick line is a ray of energy transmitted from the earth at 20 degrees north with a low take-off (elevation) angle. This ray just grazes the middle maximum and does not return through the lower ionospheric layers or earth bounce before it grazes the second southern maximum; it then bends enough to return to the earth at 30 degrees south. This layer-to-layer reflection accounts for less signal loss (10-15 dB) than the "normal" intermediate earth reflection type transmission. This low loss makes trans-equatorial propagation very effective in providing excellent DX QSOs with our southern friends.

The maximums are located on both sides of the geomagnetic equator (±20 degrees). The electrons drift-diffuse up the geomagnetic field lines each afternoon until about 2200 local time, mainly during the winter half of the year. Geomagnetic disturbances tend to enhance the electron upward drift from the trough just south of the auroral zone. The dotted lines in the drawing are representative geomagnetic field lines.

Although lower MUFs are now being experienced because of the diminishing solar cycle, one can still expect the 15- and 10-meter bands to have some trans-equatorial propagation openings this winter during the high solar flux days in which geomagnetic disturbances occur. Listen to WWV on 2.5, 5, 10, 15, or 20 MHz at 18 minutes after the hour for solar flux, geomagnetic A and K indices, as well as for the solar activity and geomagnetic condition report and forecast. If the solar flux exceeds 150 and A is greater than 30, with K greater than 4, expect very good trans-equatorial openings.

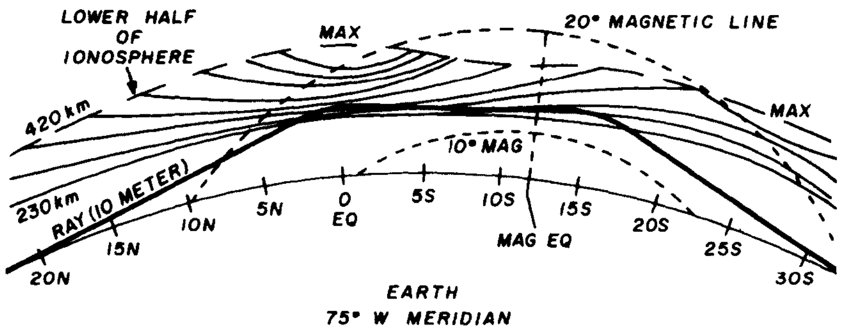


fig. 1. Trans-equatorial propagation ray path.

last-minute forecast

On the higher frequency bands (10-30 meters) DX conditions should improve with rising solar flux and flare activity during the two middle weeks of the month. This leaves the first and last weeks for operation on the improving lower bands (30-160 meters). Both night and day DX should be very good this time of the year.

Expect disturbances during the following days in October: 1st, 2nd, 9-15th, 19-21st, 24-30th. This is still the equinox period when the solar wind particles are very effective in producing ionospheric effects and still within those years of maximum solar coronal holes to feed the solar wind. All in all, that's a double dose of interesting DX conditions.

In October the Orionid meteor showers are visible from the 15th to the 25th. The maximum rate will be between ten to twenty per hour on the 20-21st of the month. The moon is full on the 21st and perigee occurs on the 4th of the month.

In most parts of the country, October represents the last opportunity of the year for antenna repair and maintenance. How are your antennas? Are

WESTERN USA

GMT	PDT	N (ASIA FAR EAST)	NE (EUROPE)	E (S. AFRICA)	SE (S. AMERICA)	S (ANTARCTICA)	SW (NEW ZEALAND)	W (OCEANIA AUSTRALIA)	NW (JAPAN)
0000	5:00	20	20	15	10	15	10	10	20
0100	6:00	20	20	15	10	15	10	10	20*
0200	7:00	20	20	15	10	15	10	10	15
0300	8:00	20	30	20	10	15	10	10	15
0400	9:00	20	30	20	10	15	10	10	20
0500	10:00	20	30	20	15	15	10	15	20
0600	11:00	*20*	*30*	20	15	20	15	15	20
0700	12:00	*20*	*30*	20	15	20	15	15	20
0800	1:00	*20*	*30*	20	20*	20	15	20*	30
0900	2:00	*20*	*30*	20	20	*20*	20	20	30
1000	3:00	*30*	*30*	*20*	*20*	*20*	20	20	*30*
1100	4:00	*30*	*30*	*20*	*20*	*20*	20	20	*30*
1200	5:00	*30*	*30*	*20*	*20*	*20*	20	20	*30*
1300	6:00	*30*	*30*	20*	*20*	*20*	20	*20*	*30*
1400	7:00	*30*	*20*	15	*20*	*20*	*20*	*20*	*30*
1500	8:00	*30*	*20*	15	20	*20*	*20*	*20*	*30*
1800	9:00	30	20	15	15	*15*	15	*20*	*30*
1700	10:00	30	20	10	15	*15*	15	15	*30*
1800	11:00	30	20	10	15	15	10	15	*20*
1900	12:00	40	20*	10	15	15	10	15	20
2000	1:00	30	20*	10	10	15	10	10	20
2100	2:00	30	15	10	10	15	10	10	20
2200	3:00	20	20*	10	10	15	10	10	20
2300	4:00	20	20	15	10	15	10	10	20

OCTOBER

MID USA

MDT	N (ASIA FAR EAST)	NE (EUROPE)	E (S. AFRICA)	SE (S. AMERICA)	S (ANTARCTICA)	SW (NEW ZEALAND)	W (OCEANIA AUSTRALIA)	NW (JAPAN)	CDT
6:00	30	20	15	10	15	10	10	20	7:00
7:00	30	20	15	10	15	10	10	20	8:00
8:00	30	30	15	10	15	10	10	15	9:00
9:00	30	30	20	10	15	10	10	15	10:00
10:00	40	30	20	15	15	10	15	20	11:00
11:00	30	30	20	15	15	15	15	20	12:00
12:00	30	*30*	20	15	20	15	15	20	1:00
1:00	*40*	*30*	20	20*	20	15	20*	20	2:00
2:00	*30*	*30*	20	20	*20*	20	20	*30*	3:00
3:00	*30*	*30*	20	20	*20*	20	20	*30*	4:00
4:00	*20*	*30*	20	20	*20*	20	20	*30*	5:00
5:00	*20*	*30*	*20*	20	*20*	20	*20*	*30*	6:00
6:00	*20*	*30*	*20*	*20*	*20*	*20*	*20*	*30*	7:00
7:00	*20*	*30*	*15*	*20*	*20*	*20*	*20*	*30*	8:00
8:00	20	*20*	15	*20*	*20*	*20*	*20*	*30*	9:00
9:00	20	20	15	20	*20*	*20*	*20*	*30*	10:00
10:00	20	20	10	15	*15*	15	*20*	*30*	11:00
11:00	20	20*	10	15	*15*	15	15	*30*	12:00
12:00	20	20*	10	15	15	15	15	*20*	1:00
1:00	30	20*	10	10	15	10	15	*20*	2:00
2:00	20	20	10	10	15	10	10	20	3:00
3:00	20	20	10	10	15	10	10	20	4:00
4:00	20	20	10	10	15	10	10	20	5:00
5:00	30	20	15	10	15	10	10	20	6:00

EASTERN USA

EDT	N (ASIA FAR EAST)	NE (EUROPE)	E (S. AFRICA)	SE (CARIBBEAN S. AMERICA)	S (ANTARCTICA)	SW (NEW ZEALAND)	W (OCEANIA AUSTRALIA)	NW (JAPAN)
8:00	30	20	15	10	15	10	10	20
9:00	30	20	15	10	15	10	10	15
10:00	30	30	15	10	15	10	10	15
11:00	30	30	20	10	15	10	15*	20
12:00	40*	30	20	15	15	15	15	20
1:00	30	30	20	15	15	15	15	20
2:00	*30*	30	20	15	20	15	15	30*
3:00	*40*	*30*	20	15	20	15	20	*30*
4:00	*30*	*40*	20	20	*20*	20	*20*	*30*
5:00	*30*	*40*	20	20	*20*	20	*20*	*30*
6:00	*20*	*40*	*20*	20	*20*	*20*	*20*	*30*
7:00	*20*	*30*	*20*	*20*	*20*	*20*	*20*	*30*
8:00	*20*	*30*	*15*	*20*	*20*	*20*	*20*	*30*
9:00	*20*	*30*	*15*	*20*	*20*	*20*	*20*	*30*
10:00	20	20	15	20	*20*	*20*	*20*	*30*
11:00	20	20*	15*	15	*15*	*20*	*20*	*30*
12:00	20	20*	10	15	*15*	15	*20*	*30*
1:00	20	20*	10	15	15	15	15	*30*
2:00	20	20	10	15	15	15	15	*20*
3:00	*20*	20	10	10	15	10	15	*20*
4:00	*20*	20	10	10	15	10	10	20
5:00	30	20	10	10	15	10	10	20
6:00	30	20	10	10	15	10	10	20
7:00	30	20	15	10	15	10	10	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.
*Look at next higher band for possible openings.

they in sufficiently good condition to last through winter? Do they need to be custom-tailored to your specific DX operating goals? Now's the time to check.

band-by-band summary

Ten to thirty meters will be open for the entire twenty-four hour period on most days of the month. The bands should peak in all directions just after local sunrise, and again toward the east and south during late evening hours. During darkness, the bands will peak toward the west, in a south-west-northwest arc that will encompass the Pacific area.

Forty and eighty meters will be the most usable nighttime DX bands. Most areas of the world will be workable from dusk until sunrise. Hops shorten on these bands to about 2500 miles for 40 meters and to 1500 miles for 80 meters, but the number of hops can increase since signal absorption in the ionosphere's D region is low during the night. The path follows the direction of darkness across the earth, similar to the way the higher bands follow the sun. Vertical antennas over good ground systems give the lowest take-off angles for long skip on these bands during darkness.

One-sixty meters will be similar to 80 meters, providing good working conditions for enthusiastic DXers who like to work the night and early morning hours. In terms of QRN, this band will be quieter now.

ham radio

product REVIEW

Cushcraft 32-19: 19-element 2-meter Boomer

If you've been paying attention to the results of VHF/UHF antenna gain contests, you'll know that the Cushcraft Boomer has been a pretty successful design. In a few instances, Boomers have been able to walk away with top honors for each band in which they were entered.

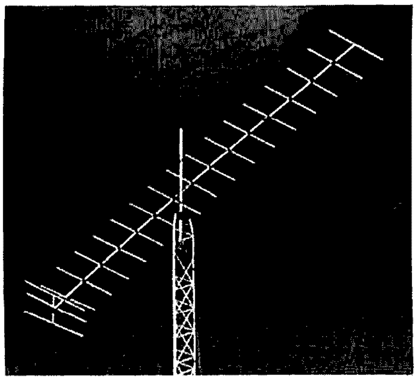
Using the latest in computer design techniques, Cushcraft engineers optimized spacing and element length to ensure maximum performance. (See **table 1**.) The Boomer also uses a T-match driven element with a 1:1 coaxial balun to provide coaxial decoupling and a low VSWR. One unique new feature on the Boomer is the Trigon three-element reflector assembly. The Trigon design enhances the front-to-back ratio and gives the antenna a sharper forward pattern.

table 1. Cushcraft 19 element 2-meter Boomer.

range	144-146 MHz
forward gain	16.2 dBd
front/back	24 dB
weight	12 pounds (5.44 kg)
length	22 feet (6.71 m)
turning radius	11 feet (3.5 m)
wind load	3.5 square feet (0.33 square m)

Assembling this antenna is a very easy job taking less than two hours to complete. All of the elements are measured and precut for ease of assembly. Fasteners are stainless steel to eliminate the problems caused by corrosion. Since the Model 32-19 uses such a long boom (22 feet) Cushcraft has added a supporting truss to provide an extra measure of mechanical stability and to keep the antenna level. All of the aluminum used is a special heat-treated tubing that is designed to take the strongest of storms without damage and keep on performing.

As many of you know, my interest in ham radio has been, for many years, 160 meters. But using the 2-meter Boomer during this June's VHF contest was a real joy. I couldn't help comparing operation on the two different bands; as I carefully and patiently pieced together the calls of stations I worked (there was fading and other propagation anomalies) I felt that my time spent ferreting calls out of the



muck on 160 meters had been excellent preparation for two-meter contesting. It was a relief not to have to fight any thunderstorm QRN. I had a lot of fun during the contest and was able to take advantage of the excellent Aurora Sunday night, June 12. Being limited to just 10 watts hurt; plans are now underway to get at least 100 watts cooking in time for the next contest. For information, contact Cushcraft, P.O. Box 4680, Manchester, New Hampshire 03108. RS#301

N1ACH

the Snyder antenna: model FB 75/80

The best way to describe this 80-meter broadband wire antenna is *simple*. For those who want to operate anywhere between 3.500 and 4.000 MHz without having to adjust wire lengths or a tuner, this 80-meter version of the Snyder product line is an answer.

From the moment of unpacking to final installation, less than 15 minutes is involved. (This assumes that a skyhook at 45 or more feet is available and the coaxial cable is already dressed.) At my location, a heavily wooded area of New Hampshire (for those of you who don't live here, it's *all* heavily wooded), a tree limb at 45 foot height was used as the support. However, telescoping masts, towers, tall poles can all serve that same function.

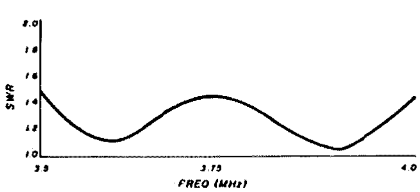
How does the antenna achieve broadband operation without tuning? It uses a patented technique that achieves this performance by combining a sealed center unit that weighs about one pound (0.45 kg) and contains no reactive elements, according to Snyder, with a middle cable assembly (that looks like a sleeve) and outer lengths of copperweld wire.

installation

Installation proceeds smoothly if you follow the instructions. In my installation I attached the supporting rope to the convenient small bracket on the center assembly and raised it five feet off the ground. After the coaxial cable was connected and sealed, and stess relieved (45 feet of RG-8 weighs a bit), I unfurled the middle cable assemblies and payed them out. I attached ropes to the six-inch long end insulators and raised the antenna to its final height while securing the copperweld ends, which I left for last. (Those of us who have worked with copperweld – haven't we all? – can appreciate the small additional amount of time and care required in handling this spring wire. *Unwind only the section of cable that you're working with.*)

The FB 75/80 makes an attractive, sturdy installation as an inverted vee. It can be secured in other configurations as a horizontal dipole, sloping dipole, or as an element in an array. It is highly recommended that if used as a horizontal dipole a center support be provided to bear the weight of the feedline and center insulator.

SNYDER MODEL FB-75/80 ANTENNA



performance

This antenna has an SWR curve that has the shape of the letter *W* (see fig. 1). Feeding a small amount of power into this antenna at every 25 kHz from 3.5 to 4.0 MHz and using a Bird thruline wattmeter basically confirmed the manufacturer's specifications. This antenna "wants" to be in the clear away from all large conducting and dielectric objects or obstructions to provide best performance.

On the air comparisons between this antenna in an inverted-vee configuration and a 50-foot apex high sloping delta loop revealed the expected results. Using a rapid switching arrangement it compared favorably (same report) with the Delta loop antenna for most U.S. contacts. The sloping delta loop has the edge for Europe and longer paths. In all fairness it must be mentioned that the sloping delta loop performs as well as my full size 80-meter Bobtail curtain in its favored directions.

conclusion

The Snyder Antenna Company has accomplished its goal in providing Amateurs with a non-tune broadband basic antenna. It is a pleasure to install and operate with any transceiver and especially complements any of the modern "instantaneous" frequency change solid-state units. In addition to base station use, it is well suited for portable operation be It emergency, field day, or DXpedition. For information about Synder products, contact the company at 250 East 17th Street, Costa Mesa, California 92627. RS#303

K2RR

BBC Metrawatt model MA3E multimeter

There's no doubt about it: digital readouts are definitely in vogue. So it came as some surprise when an analog multimeter appeared on my desk for review the other day.

BBC Metrawatt entered the American market fairly recently when it introduced a rather complete line of both digital and analog meters. Model MA3E represents a moderately-priced unit that offers a number of appealing features.

My first impression was that this is a very European unit, with its gracefully curved lines in a very neat and functional design. (No wonder — its unique case was designed by Porsche.) To protect the meter when not in use, the unit folds together to make a convenient, easy-to-transport package. The high-impact plastic case is built to withstand industrial use or abuse without compromise in performance.

The MA3E incorporates an electronic amplifier providing a combined total of 35 measuring ranges that include 18 for AC and DC voltages up to 1000 volts; 12 for AC and DC currents up to 10 amperes; and five resistances up to 20 megohms, including a battery check — all while providing a constant input resistance of 10 megohms.

All measuring ranges are manually selected by use of a switch located on the main frame. The viewing angle of the meter can be varied to meet the needs of the user.

The unit operates from an internal 9-volt battery or from an external power supply. Low power consumption ensures long battery life.

The test leads are recessed into the meter and specially shielded to prevent accidental shorting while plugged into the meter. The leads are also threaded to accept alligator clips and other accessories.

The MA3E measures approximately 5-3/4 × 4-1/2 × 1-3/4 inches (146 × 118 × 44 mm) and weighs less than one pound (0.45 kg), without battery.

Luckily the MA3E arrived for review while we were rebuilding WB1AHV, the *ham radio* Amateur station. While crawling up and down connecting power lines and antenna leads, I found the MA3E to be a real joy to use. The adjustable meter made reading voltages or determining continuity a snap. Being a bit ham-fisted, I accidentally knocked the meter off the operating table, but the rugged case wasn't damaged at all, to my relief.

This is a useful meter that, at under $200, should find wide acceptance in both Amateur Radio and industrial applications.

Detailed information on the MA3E and other models in the product line are available from BBC-Metrawatt, 6901 West 117th Avenue, Broomfield, Colorado 80020. RS#302

N1ACH

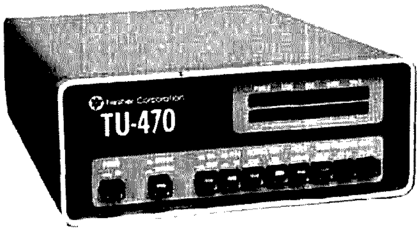
NEW products

RTTY/CW terminal

The new Flesher Corp. TU-470 RTTY/CW terminal unit receives up to 300 baud on all three shifts and provides TTL and RS-232 compatible I/O, including bi-polar CW and PTT outputs, for complete remote control and isolation of computer level I/O keying.

Each TU-470 RTTY filter board is a high-sensitivity, high Q, 3 stage, 6 pole active bandpass filter said to provide excellent stability and sharpness. A signal balance restorer circuit has been incorporated to allow reception of non-standard RTTY shifts on mark only. The CW filter/demodulator has a 3 stage, 6 pole filter centered at 750 Hz for CW reception.

The unit also provides crystal controlled AFSK, FSK, 170 Hz narrow preselector filter, built-in 20 or 60 mA loop supply, autostart, threshold control, 5 LED indicators, bar graph tuning, scope outputs, reverse receive, reverse transmit, and much more.



The TU-470 RTTY/CW terminal unit is available wired and tested for $499.95.

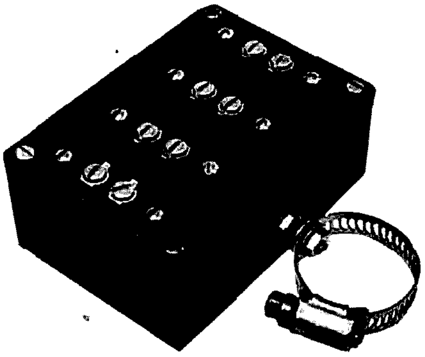
For more information, contact Flesher Corporation, P.O. Box 976, Topeka, Kansas 66601. RS#304

lightning protector

The new IS-RCT from PolyPhaser relies on fast 50 nanosecond three-element crowbar gas tubes to provide effective protection from lightning for the hamshack rotor control box.

Designed to be mounted on a ground pipe or grounded tower leg, with terminals down to keep snow and rain off connectors, the unit is housed in a heavy duty stabilized plastic case. All grounding hardware is stainless steel.

Crimp and/or solder lugs are included. The introductory price is $29.95.



For information, contact PolyPhaser Corporation, 1500 West Wind Boulevard, Kissimmee, Florida 32741. RS#305

keyer trainer

The AEA Basic Trainer Model KT-3 is a computerized Morse code instructor. Character speeds are given at 20 WPM with a three-second interval between letters or letter groups. This method of learning helps eliminate the plateau so many Morse code students encounter between 10 and 12 WPM. This system does not encourage the student to learn a "dot-dot-dash-dot" for F, but rather one cohesive sound of dididahdit (at the character speed of 20 WPM). As a result, the student does not have to unlearn bad habits such as counting dots and dashes in order to copy code.

Because the sound of any given letter does not change from 0 to 20 WPM, it is easier for the student to increase copying speed if he or she has learned code at a rate of 20 WPM. To facilitate this learning technique, the minimum programmable speed on the Basic Trainer is 18 WPM character speed. After learning the code, the user can progress up to 99 WPM with the KT-3. Each character is taught separately by repetition. The student progresses only after he is confident he knows the letters being presented by the KT-3.

After the first letter (F, in this case) is learned, a student may progress to the letter K. After learning K, the student activates the computer to present the letters F and K in random sequence at a 20 WPM character speed. This technique continues as the student learns each subsequent letter in the alphabet.

Easy to operate, the KT-3 requires no computer programming skills. An earphone monitor jack is provided for private practice sessions. The KT-3 operates from 12 Vdc (or 117 Vac with optional AC-1 Wall Adaptor).

For details, contact Advanced Electronic Applications, Inc., 2006-196th St. SW, Lynwood, Washington 98036-0918. RS#306

six-band high-frequency vertical

Hustler has announced the availability of its new 6-BTV six-band trap vertical for the high-frequency ham bands.

Based on the popular 4-BTV, the new 6-BTV offers full band coverage of 10, 15, 20, 30, and 40 meters with a VSWR of under 1.6:1 at band edges and up to 100 kHz on 75 or 80 meters.

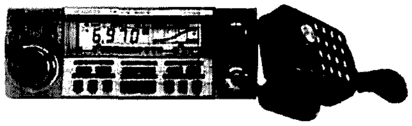
According to the manufacturer, fiberglass reinforced high *Q* traps, extra-strength aluminum components, and stainless steel hardware combine to make the new 6-BTV the most rugged, high performance, high-frequency vertical available.

The 6-BTV lists for $199.95. For additional information, contact Hustler, Inc., 3275 North B Avenue, Kissimmee, Florida 32741. RS#307

fm dual band

Trio-Kenwood has announced the release of a compact new combination 2-meter and 70-centimeter FM radio, model TW-4000A. Among its features are a large, easy-to-read LCD display, ten channels of memory with offset recall, lithium battery memory back-up, dual digital VFO's, priority watch, common channel, programmable memory scan, band scan, and a full 25 watts of rf output on each band.

An optional accessory available for use with the TW-4000A is the VS-1 voice synthesizer unit that announces the operating frequency,



VFO "A" or "B", repeater offset, and the memory channel number when the unit is turned "ON", when another frequency is selected, or when a memory is recalled. The VS-1 is designed to be easily installed inside the TW-4000A.

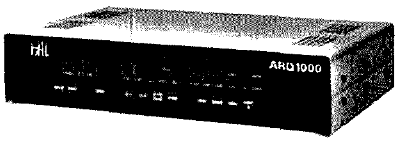
Additional information is available from Trio-Kenwood Communications, 1111 West Walnut Street, Compton, California 90220.

AMTOR terminal

The ARQ1000 is a full send-receive terminal for the AMTOR ARQ code. All features of the CCIR 476-2 Recommendation are supported. Modes include: ARQ, FEC, SEL-FEC, and MONITOR. The ARQ1000 may be used with the HAL DS3100 and ST6000, CT2200, CT2100, or CRW6850 terminals or any ASCII or Baudot terminal at baud rates from 45 to 300 baud. Non-volatile keyboard-programmable

NEW products

ARQ access code, SEL-CAL code, and WRU answer-back codes are included. The ARQ1000 is housed in a cabinet that matches the CT2200 and CT2100. Available options include the DM170 internal demodulator and ARQX10 *encryptions module.*



For more information, contact HAL Communications Corporation, P.O. Box 365, Urbana, Illinois 61801. RS#311

single chip code-practice processor

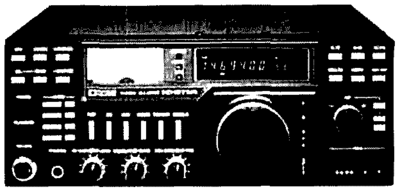
The world's first single chip code-practice processor, the CPP1, is a true 8-bit microcomputer containing all the tables and timing necessary to learn Morse code. By adding a simple dot clock and tone generator to this processor, it is possible to practice Morse code at rates from 1 WPM to over 100 WPM.

The CPP1 is said to eliminate the frustrations of trying to learn code by tape, and promises to put an end to searching for a particular practice group, stopping and rewinding, and having to learn at whatever speeds the tape offers.

For further information, contact Micro Digital Technology, P.O. Box 1139, Mesa, Arizona 85201. RS#309

base station transceiver

ICOM has announced the introduction of a new base station transceiver for 2 meters, the IC-271A. Covering the entire 2 meter ham band, it features fm/upper sideband/lower sideband and CW modes, has a 25 watt output standard, with an optional built-in power sup-

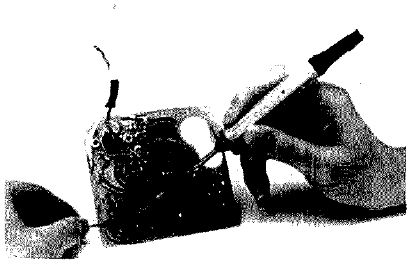


ply available. It has 32 full function memories. Built-in sub-audible tones selectable from the main tuning dial provide ease of operation. Frequencies, modes, tones and offset may be written into each memory. Scanning is possible with the IC-271A; either the whole band, memories or selected modes may be scanned. The IC-271A features ICOM's new high contrast, two-color display, showing frequency digits in white and control functions in red.

For further information, contact ICOM America, Inc., 2112-116th Ave., NE, Bellevue, Washington 98004. RS#310

miniature soldering iron

An industrial grade precision miniature soldering iron that heats up quickly, has a non-



melt handle and is suited for students, hobbyists, and professionals is available from M.M. Newman Corporation.

The Antex Model G Soldering Iron features a wide range of slide-on tips. Designed for delicate circuitry, the tips are directly grounded and heat up in only 45 seconds. Compact and fully portable, the Antex Model G Soldering Iron weighs only 3/4 ounce without cord. Built for continuous or intermittent use, the tip slides directly over the heating element and heats up to 700-750 degrees F. Complete with a premium grade pretinned 3/32 inch tip, the Model G sells for $15.95.

For more information, contact M.M. Newman Corporation, 7 Hawkes Street, P.O. Box 615A, Marblehead, Massachusetts 01945. RS#312

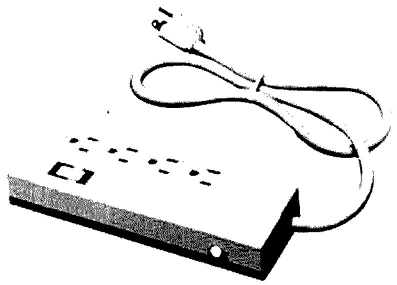
multi-outlet panel

A new multi-outlet panel from Flexiduct provides four circuit-breaker-protected outlets for safe use where too few electrical outlets exist.

The new outlet-extender panel features 4 grounded outlets, an on/off switch, a circuit breaker with a push-button reset (to protect from overloads) and a double-insulated cord with a three-prong end plug. It is rated at 15 amps, 125 volts, 1875 watts and is UL listed.

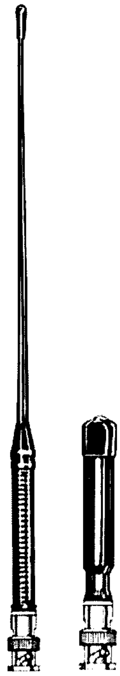
Other Flexiduct safety products include 4 and 6 outlet multi-outlet strips, a plug-in surge suppressor, a multi-outlet surge suppressor,

lay-flat power extensions, over-the-floor cord cover and products for use with telephones.



Information on Flexiduct safety products is available from Winders & Geist, Inc., P.O. 83088, Lincoln, Nebraska 68501. RS#313

two more Kulduckies



Larsen Electronics has added two new antennas to its Kulduckie line. Both the Helical Quarter Wave and the Stout are designed to work with most hand-held radios.

The Helical Quarter Wave antenna combines the best features of the helical and full quarter wave antenna. It is said to deliver better performance with a shorter, more stable antenna than a full quarter wave VHF. The Stout antenna series features a slightly larger diameter helix with a closer pitch for applications in which space is a problem. Both new antennas are available in VHF frequencies 136-142, 142-150, 150-162, and 162-174 MHz.

For more information, contact Larsen Electronics, P.O. Box 1799, Vancouver, Washington 98668. RS#314

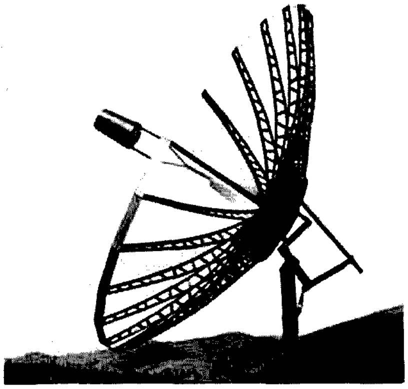
new dish

A satellite TV antenna made by an entirely new method of production has been announced by Total Television, Inc.

The design of the 12 foot diameter dish consists of a heavy-duty expanded aluminum mesh reflective surface supported by twenty-four injection molded ribs connected to an injection molded main support.

The material from which the dish structure is molded is a sophisticated plastic formulated to endure all kinds of weather. Products molded from the plastic material have been used for nearly twenty years to replace concrete in exterior and underground applications. The manu-

NEW products



facturer says that the material, Structuralite,™ will be unaffected by temperatures ranging from minus 55 degrees to plus 255 degrees F. grees F.

The mesh reflective surface is suitable for 4 GHz microwave signals used in all current U.S. domestic satellites, yet will allow the wind to pass and permits the background to be seen through it.

For further information, contact Total Television, Inc., 17537 N. Umpqua Highway, Roseburg, Oregon 97470. RS#315

dipole

The new KT5B multi-band dipole antenna from Kilo-Tec is designed for use on the 1.8 MHz through 30 MHz amateur radio bands. It uses no loading coils or traps and will handle 2 kW PEP. The price is $59.95.

For further information, contact Kilo-Tec, P.O. Box 1001, Oak View, California 93022. RS#316

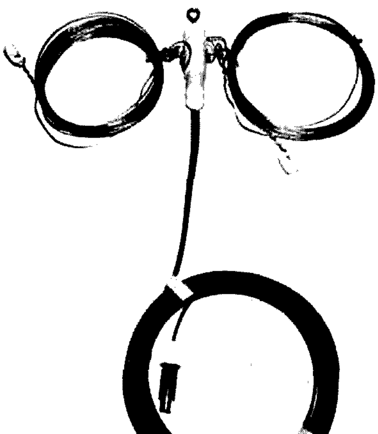


plate frames

License plate frames personalized with individual call signs are now available from BHC, Incorporated. Molded from black ABS plastic — the same material used for trim on many new cars, a set of HAM-TAGS consists of two

black frames with white, permanent vinyl letters in the large imprint area. (License plates differ from state to state so drivers would have to check their plates to see if their calls would go at the top or bottom of the frame.) In states that issue only one plate, BHC will furnish a frame for the rear and a plate for the front. HAM-TAGS are $12.95 per set plus $1.50 shipping; information is available from BHC, Incorporated, 1716 Woodhead, Houston, Texas 77019. RS#317

scaleable ac voltmeter

The Slimline II ac voltmeter from Nationwide Electronic Systems, Inc., (NES), scales ac voltage input signal to any desired unit, mounts quickly, and features a large (0.9 inch) LED display which can be easily scaled to read out directly in engineering units. The Slimline will accept the output from a voltage transformer, ac tach generator, or any device with an ac voltage output.

Examples of use include scaling 34.52 Vac from a potential transformer to read directly as 1.726 kV; 0.825 Vac from a current-to-voltage transformer to read directly as 165 ac amps; and 43.5 Vac from ac tach generator to read directly as 783 rpm.

All adjustments and controls for this meter are accessible under the flip-up door located beneath the display. Models are available for ranges from 0 to 200 Vac.

Also available from NES are scaleable dc voltmeters, scaleable ammeters, clocks, counters, process meters, and the unique ASCII Bustle* (BCD/ASCII converter). All NES products are backed by a solid 3-year warranty.

For more information or a copy of the NES Condensed Catalog, contact NES, Inc., 1536 Brandy Parkway, Streamwood, Illinois 60103. RS#318

field-strength element

An extremely sensitive relative field-strength element, Model 4030, expands the usefulness of the thousands of Thruline wattmeters in the field by helping to optimize the radiated signal of any transmitter from 2 to 1000 MHz.

Model 4030 employs modern broad-band circuitry instead of the highly reactive resonant networks of most field-strength meters, which limit their utility. The element consists of a flexible receiving antenna, a single high-pass network and a variable gain rf amplifier/detector. A battery-saving feature turns everything off when the element is removed from the wattmeter. Typically full scale deflection is obtained from a one watt CW source at 150 MHz through a quarter wave antenna 8 feet distant. Dynamic range is at least 30 dB and battery life is 100 hours or more.

For more information, contact Bird Electronic Corporation, 30303 Aurora Road, Cleveland (Solon), Ohio 44139. RS#319

✓ 149

"affordable" repeaters

A new line of repeaters in several frequency ranges is available from Internatinal Telecommunications Systems Florida Inc. Ranges include 30-50 MHz, 132-174 MHz, 200-240 MHz, 380-480 MHz, or any band combination. The power output is 30 watts; the unit operates from 13.6 VDC or 115/220 Vac and costs $499.

For further information, contact International Telecommunications Systems Florida Inc., 8416 N.W. 61st Street, Miami, Florida 33166. RS#320

NEW products

handheld DMM

Keithley Instruments, Inc., has introduced the Model 132 Handheld DMM as their top of the line 3-1/2 digit meter. The unit combines the field service capabilities of Keithley Handheld DMM's with the two most often required additional measurement capabilities, TRMS ac and temperature.

The 132 features unique characteristics: according to Keithley, it is the only handheld DMM which incorporates TRMS and TEMP; carries the lowest price of any handheld TRMS DMM available; and is the only one with an integral TC connector and full probe selection.

Available in both a degree F version (132F) and a degree C version (132C), the 132 has a complete dc voltage range from 200 mV to 1000V with 0.25 percent accuracy, current ranges from 2 mA to 2 A and resistance ranges from 200 ohms to 20 megohms. The 132C measures temperature from −20 degrees to 1370 degrees C, the 132F from 0 degrees F to 2000 degrees F using optional type K thermocouple sensors or probes.

For complete information, contact Keithley Instruments, Inc., 28775 Aurora Road, Cleveland, Ohio 44139. RS#321

fm transceivers

Amateur-Wholesale Electronics has announced the new PCS-4000 series of FM transceivers. The familiar PCS-4000 covers the 2-meter band plus adjacent CAP and MARS frequencies, from 142 to 149.995 MHz, with 25 watts output. Three new units are now available for 10 meters, 6 meters, and 70 centime-

ters. The PCS-4800 covers 28 to 29.99 MHz in steps of 10 or 20 kHz. The PCS-4500 covers 50 to 53.995 MHz in steps of 5 or 10 kHz. The PCS-4300 covers 440 to 449.995 MHz in steps of 5 or 25 kHz. A 220-MHz unit will soon be available.

The new PCS-4800, PCS-4500, and PCS-4300 employ the same advanced microcomputer chip found in the PCS-4000, providing the versatility and ease of operation found in the 2-meter Azden transceiver. Standard repeater splits, 16 memory channels in two separate banks, up to eight odd splits, dual memory scan, dual programmable band scan, free and vacant scanning modes, and frequency reverse are but some of the features. Each unit has 10 watts or 2 watt selectable output, discriminator scan centering, a digital S/RF meter, busy-channel and transmit LED indicators, and a full 16-key autopatch pad (except for the PCS-4800).

All units come with dynamic remote-function microphone, built-in speaker, mobile mounting bracket, remote speaker jack, and all hardware. The size is 2 inches high by 5-1/2 inches wide by 6-3/4 inches deep. The bright green LED frequency display and illuminated keyboard provide visibility under a variety of ambient conditions. All are designed for mobile or fixed station use.

For further information, contact Amateur-Wholesale Electronics, 8817 S.W. 129 Terrace, Miami, Florida 33176. RS#322

SSTV/RTTY/CW/FAX tuner

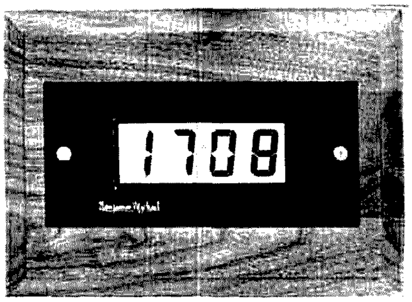
The new "Blinky" Model 959 is a SSTV tuning unit consisting of an op-amp limiter amplifier and six active precision tuned, temperature stable bandpass filters, providing a visual blinking LED indication when a received SSTV, RTTY, FAX, or CW signal is perfectly tuned. The

user simply tunes until all six LED's are blinking at the same time and the SSTV is tuned. For RTTY, FAX, or CW, the user tunes until the first two LED's are blinking (low tones) or the last two LED's (high tone).

"Blinky" Model 959 is made in the United States and sells for $99.95. For more information, contact TimeKit, P.O. Box 22277, Cleveland, Ohio 44122. RS#324

digital military time clock

The Model 193A Digital Military Time Format Wall Clock from Benjamin Michael Industries, Inc., features a 1-inch digital display for excel-



lent visibility. It also features quartz accuracy, total immunity to power line failures and disturbances, and all solid-state construction. The clock will operate for well over one year on a single AA battery.

Housed in a hand-made solid walnut case, the model 193A is priced at $129.95 plus $3.00 shipping. For more information, contact Benjamin Michael Industries, Inc., 65 East Palatine Road, Prospect Heights, Illinois 60070. RS#323

CU1239/ARC-105, CU-1669A, ARC-105, Sperry Rand 3226A1, 3226B1, 690D1, 490B-1. Top dollar paid or trade for new Amateur gear. Write or phone Bill Slep, 704-524-7519, Slep Electronics Company, Hwy. 441, Otto, NC 28763.

SUPER COWW CONTEST programs. TRS-80 Model I, III, (IV in III mode) Completely machine language. Logs 3200+ contacts per band. Automatic identification of country and zone from call letters. Dupe Speed 12000+ contacts per second. Screen displays zones still needed, total points, zones, countries, etc. Automatic cw generator with 2 buffers. Log Print program prepares logs and dupe sheets. $39.95 Log Preparation program for hand logs. Similar features to above. $29.95 Free fact sheet and sample printouts. K4SB, 3496 Velma Dr., Powder Springs, Georgia 30073.

Coming Events

ACTIVITIES

"Places to go..."

CONNECTICUT: The Tri-City ARC Auction, St. James Parish Hall, Poquetanuck, October 29. Setup 9 AM, auction 10 AM until sold out. Admission free. Food available. Talk in on 146.73, 67 or 94 repeaters. For information call WA2RYV (203) 848-9670.

GEORGIA: The Alford Memorial Radio Club's 11th annual Hamvention, November 5 and 6, Stone Mountain Park, Stone Mountain. 9 AM to 5 PM Saturday and 9 AM to 3 PM Sunday. FCC exams, seminars, dealers, gigantic flea market, free parking. $3.00 admission fee includes Saturday night family cookout and variety entertainment. Camping and other amusements at the park. Talk in on 146.16/146.76. For further details: SASE to Lew Howard, W4LHH, 4132 Creek Stone Court, Stone Mountain, GA 30083 or (404) 292-5469.

INDIANA: The Allen County Amateur Radio Technical Society will hold their 11th Fort Wayne Hamfest, Allen County Memorial Coliseum, November 13. OSCAR, FSTV/SSTV, Audio and computer forums. Tickets $2.50 advance (SASE). $3.00 at door. Children under 12 free. Parking $1.00. Deadline for advance tickets November 1. Tables $6.00. Premium tables $20.00. Talk in .88 and .52. For information: Hamfest Chairman, AC-ARTS, Inc., P.O. Box 10342, Fort Wayne, IN 46851.

INDIANA: The Hoosier Hills Ham Club's 22nd annual Hamfest, Sunday, October 9, Lawrence County 4-H Fairgrounds, U.S. 50, Bedford. Admission $3.00; swap shop $2.50 with own tables. Gates open 10 AM Saturday, October 8, for flea market setup and campers. Free fish fry, campfire, entertainment, coffee, Saturday night. Talk in on 146.13-73 on 3910 kHz. For information: KA9JTZ, Dick Reistter, Hoosier Hills Ham Club, Box 891, Bedford, Indiana 47421.

LOUISIANA: The Twin City Ham Club's Hamfest, Saturday, November 12, West Monroe Convention Center, 8 AM to 5 PM. Delaers — Swap Tables — Fellowship. Talk in 25-85 and 52-52. For information: AE5V, Benson Scott, 107 Contempo, West Monroe, LA 71291.

LOUISIANA: Amacom '83, the New Orleans Hamfest-Computerfest, October 15 and 16, Delgado Community College's City Park campus. Sponsored by the Jefferson ARC, Greater New Orleans ARC, the Delta DX Association and the New Orleans VHF Club. Demonstrations, speakers, awards, flea market and commercial exhibits. FCC exams. Admission $5 advance (incl. two tickets) or $5 at door (no tickets). Talk in on Jefferson ARC's repeater W5GAD/R on 147.285/.885 or 449.0/444.0. For details and reservations: Amacom '83, P.O. Box 73665, Metairie, LA 70033 or call Bill Bushnell, WA5MJM, Chairman, at (504) 887-5022. Reservations should be made before October 5.

MARYLAND: The Columbia Amateur Radio Association's 7th annual Hamfest, Sunday, October 23, 8 AM to 3:30 PM, Howard County Fairgrounds, 15 miles west of Baltimore. Admission $3.00. Indoor tailgating and tables $6.00 additional. Outdoor tailgating $3.00 additional. Talk in on 147.735/135, 146.52/52. For reservations and information: Ed Wallace, K3EF, 9905 Carillon Drive, Ellicott City, MD 21043.

MASSACHUSETTS: The Framingham Amateur Radio Association's 9th annual Fall Flea Market, the largest indoor flea market in New England, Sunday, October 30. We will again be at the Framingham Civic League Building, 214 Concord Street (Route 126) in downtown Framingham. Doors open 10 AM; sellers setup 8:30 AM. Admission $2.00. Tables $10.00 (preregistration required). Radio equipment, computer gear, bargains galore! Talk in on 75/15 and 52 direct. For information/reservations: Ron Egalka, K1YHM, 3 Driscoll Drive, Framingham, MA 01701.

MASSACHUSETTS: The Wellesley Amateur Radio Society's annual Fall tailgate flea market, Wellesley High School, Rice Street, Wellesley. Starts 9 AM. Admission $1.00 for buyers or sellers. Talk in on 147.63/03. For information: Nels Anderson, K1UR, (617) 872-5259.

MASSACHUSETTS: The 1983 New England DXCC Dinner, November 5, Concord Lodge of Elks, Baker Avenue, West Concord. DX talk and slide programs, auction begins at 2 PM. Afternoon session $2.00. Cocktail hour 6 PM followed by a 7-course roast beef dinner. Banquet speaker VK0CW. Cost for the evening $14.95. Reservation forms will be mailed in September. Contact Jim Dionne, K1MEK, 31 DeMarco Road, Sudbury, MA 01776.

MICHIGAN: The Central Michigan Amateur Radio Club and Lansing Civil Defense Repeater Association will hold their annual Ham Fair, Sunday, October 9, 8 AM to 3 PM, Grand Ledge High School 7 miles west of Lansing. Swap shop, demonstrations, antennas, computers and handcrafted items. Donation $2.50 adults. Tables 75¢/foot. For information: Rowena Elrod, KA8OBS, 111 Lancelot Place, Lansing, MI 48906, (517) 372-5462; or write Ham Fair '83, P.O. Box 18044, Lansing, MI 48901.

MICHIGAN: Electronic Flea Market, Sunday, October 23, Kalamazoo County Fairgrounds, 10 AM to 4 PM. Dealer setup 9:30 AM. Tickets $2.00 advance; $2.50 at door. 4 ft. table rental $2.50 advance; $3.00 at door. For table reservations and tickets SASE with check to: Ham 10 FM Club of Kazoo, Ken Cosey, KA8RUA, 2825 Lake Street, Kalamazoo, MI 49001.

NEW ENGLAND: The Hosstraders will hold their annual autumn swapfest, Saturday, October 8, rain or shine, Deerfield, New Hampshire, Fairgrounds. Admission $1.00, tailgating included. Friday night camping for self-contained rigs, after 4 PM. Profits benefit Shriners' Boston Burns Hospital. Last May's donation $2,702.00. For information/map SASE to Norm, WA1IVB, RFD Box 57, West Baldwin, ME 04091; or Joe, K1RQG, Star Rt. Box 56, Bucksport, ME 04416; or Bob, W1GWU, 105 Walton Rd., Seabrook, NH 03874.

OHIO: The Northwest Ohio Amateur Radio Club's 9th annual Lima Hamfest, Sunday, October 9, Allen County Fairgrounds (exit 125/126, 1 mile from I-75). Gates open 6 AM. Tickets $3.00 advance; $3.50 at gate. Tables $6.00; half table $3.50. Free camping, electrical hookup $7.00. For information/reservations (please enclose check): N.O.A.R.C., Box 211, Lima, OH 45802. Talk in on 146.07/67, 147.63/03, 146.52/52.

OHIO: The Marion Amateur Radio Club will hold its 9th annual "Heart of Ohio" Ham Fiesta, Sunday, October 30, 0800 to 1600, Marion County Fairgrounds Coliseum. Tickets $3.00 advance, $4.00 at door. Tables $5.00. Check in on 146.52 or 147.90/30 or 223.34/224.94. For information, tickets or tables contact: Paul Kilzer, W8GAX, 393 Pole Lane Road, Marion, Ohio 43302. (614) 389-5573.

PENNSYLVANIA: The R.F. Hill Amateur Radio Club's annual "Winterfest" Amateur Radio Flea Market, Sunday, November 13, Sellersville National Guard Armory, Route 152, Sellersville. Indoor space available $4.00. Outdoor tailgating $3.00. Buyers $2.00. Refreshments on site and nearby. Flying hams land at Pennridge Airport, Perkasie. Talk in on 144.71/145.31 Almont. 146.28/146.88 Souderton and 146.52 simplex in local area. For information: P.O. Box 29, Colmar, PA.

PENNSYLVANIA: The Irwin Area ARA's Swap & Shop, Saturday, October 15, Circleville V.F.D., off U.S. 30, 3.5 miles west of Penn. Turnpike, exit 7. Flea market, vendors, refreshments, free parking. Talk in on 146.925/.325 and 146.52 MHz. For information: Rick Jackson, N3DAA, 39-D Lower Boone Drive, Turtle Creek, PA 15145. (412) 829-1953.

PENNSYLVANIA: The Red Rose Repeater Association and SERCOM, Inc., will sponsor a Computerfest, Sunday, October 23, 9 AM to 4 PM, Guernsey Sales Pavilion, east of Lancaster, Rts. 30 and 896. Admission $3.00. XYL's and children under 14 free. Featuring computers and Amateur Radio equipment. Tailgating $2.00. Inside tables available by reservation. Talk in on 147.615/015, 146.01/61 and 146.52 simplex. For information: The Computerfest Committee, P.O. Box 5029, Lancaster, PA 17601.

TENNESSEE: Hamfest Chattanooga and the State ARRL Convention, October 22 and 23, Chattanooga State Technical Community College, Amnicola Highway, Chattanooga. Forums, contests and non-ham activities. College cafeteria will be serving breakfast and lunch both days. A hospitality party will be held Saturday, October 22, at the Ramada Inn. Phone (615) 894-6110 for reservations and ask for special "Hamfest Chattanooga" rates. Inside dealer and indoor/outdoor flea market spaces available. For information: Hamfest Chattanooga, P.O. Box 3377, Chattanooga, TN 37404 or phone N4DON (404) 820-2065.

AMSAT will hold an Amateur Radio Satellite Symposium in conjunction with its annual meeting, Saturday, November 12, Johns Hopkins University Applied Physics Laboratory, off I-95 between Baltimore and Washington. Program includes tracking OSCAR 10, a report on the W5LFL space shuttle operation, PACSAT and more. Free admission but reservations are required. Contact: AMSAT, P.O. Box 27, Washington, DC 20044 or (301) 589-6062.

OPERATING EVENTS

"Things to do..."

OCTOBER 1: The Jackson (TN) Amateur Radio Club will operate special events station K4EP from the historic Casey Jones' Railroad Museum to help celebrate heritage week in Jackson. Operation will be on the lower 25 kHz of the General class phone bands on 80, 40, 20 and 15 meters. All QSL's will be answered with a unique commemorative card. Please SASE. Hams wishing to visit the home of the famous railroader use 147.21 up 600 for information and directions.

OCTOBER 8 AND 9: Columbus Day Special Event. Open to Radio Amateurs worldwide. Columbus hams work non-Columbus. Non-Columbus hams work as many Columbus hams as required for certificate. Saturday, October 8, 1400Z to 2400Z, 10 meters phone, 28.6 MHz ± 10 kHz. Sunday, October 9, 1400Z to 2400Z, 15 meters phone 21.4 MHz ± 10 kHz. Exchange name and RST. Score one point for each contact (excluding W8TO). Six points for a W8TO contact. A final score of ten must be submitted within 120 days to be eligible for certificate. Please include SASE. Requests for certificates: Radio Station W8TO, Special Event Coordinator, 280 East Broad St., Columbus, Ohio 43215.

OCTOBER 8 AND 9: The Peninsula Radio Operators Society (P.R.O.S.) will operate special event station KB3QV on the grounds of Salisbury State College, Salisbury, Maryland, to help celebrate the Delmarva Folklife Festival. 1600Z to 2100Z both days. Frequencies: 10 through 80 meters General phone bands. QSL information: SASE to P.O. Box 2315, Salisbury, MD 21801.

OCTOBER 11, 12 AND 13: The Colquitt County Ham Radio Society will operate club station WD4KOW from the site of the sixth annual Sunbelt Agricultural Exposition, 0900 to 1700 EDST each day. Operations will be in the General portion of the HF bands. Members will be listening for visiting hams on local repeater 146.19/79. Visiting hams are welcome to the Amateur booth at the Expo and to operate the Amateur station. For a special QSL card, please SASE to Colquitt County Ham Radio Society, P.O. Box 813, Moultrie, GA 31768.

OCTOBER 15 AND 16: Jamboree on the Air. 0001Z Saturday to 2400Z Sunday. The 26th annual Scouting-ham radio event sponsored by the World Scout Bureau, Geneva, Switzerland. Scouts of all ages are welcome to participate. The World Bureau station, HB9S, the BSA station, K2BSA, and many others will be operating. Exchange friendship greetings, scout talk and experiences. For certificate SASE to Boy Scouts of America, International Division/JOTA cards, 1325 Walnut Hill Lane, Irving, TX 75062-1296. A special pocket patch is available for $1.00 each pp, any quantity. Send check/M.O. to BSA, International Division/JOTA patch, same address. Frequencies: CW — 3,590, 7.030, 14.070, 21.140, 28.190 MHz. Phone: 3.940, 7.290, 14.290, 21.360, 28.990 MHz plus Novice, RTTY and SSTV frequencies.

OCTOBER 22 AND 23: QRP Amateur Radio Club International Fall QSO Party. 1200 UTC Saturday, to 2400 UTC Sunday. Maximum operating time 24 hours. Exchange RS(T), state/province/country and QRP ARCI membership number. Non-members give RS(T), state/province/country and power output. Frequencies: CW — 1810, 3560, 7040, 14060, 21060, 28060, 50360 kHz. SSB — 1810, 3985, 7285, 14285, 21385, 28885, 50385 kHz. Novices and Technicians: 3710, 7110, 21110, 28110. Call CQ QRP de (call sign) CQ QRP Contest etc. Certificates awarded to highest scoring station in each state/province/country with two or more entries. All entries considered for Triple Crowns of QRP award. Send full log data plus separate worksheet showing details, time off the air. None returned. For results and scores include large SASE and one ounce of U.S. postage or IRCs. Logs must be received by November 20, 1983. Send to QRP ARCI Contest Chairman, William Dickerson, WA2JOC, 230 Mill Street, Danville, PA 17821.

OCTOBER 28: Suffolk County Radio Club will operate W2DQ from 0000Z, October 28 to 2400Z October 30 to celebrate Suffolk County's 300th birthday. Frequencies: Phone 15 kHz up from lower 40-15 meter General class band edges. Novice 21.135. For a special certificate send large SASE to AC2P.

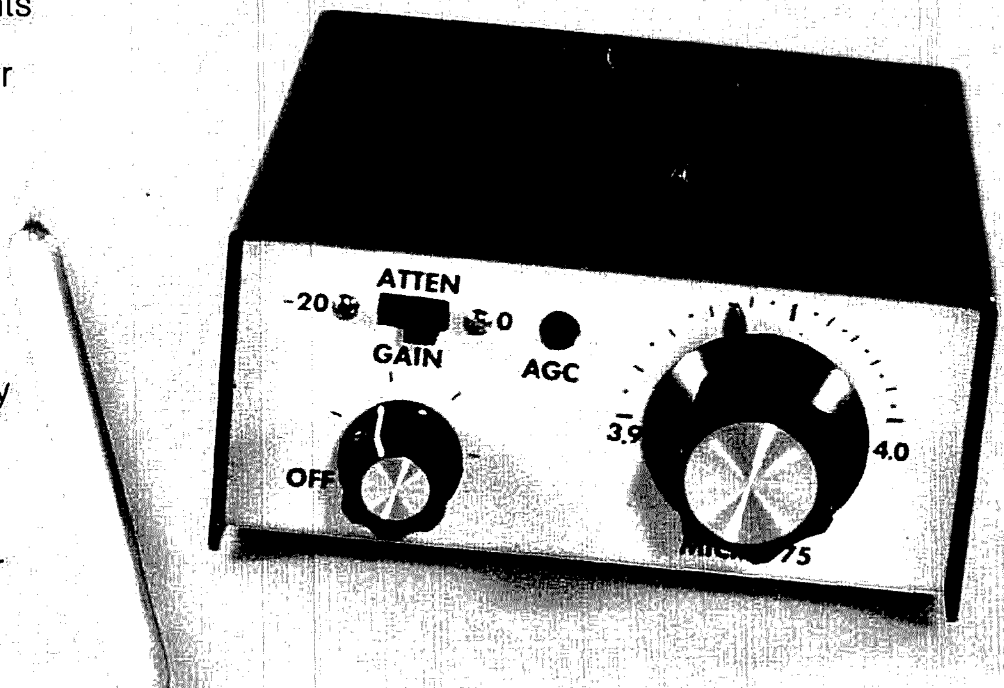
OCTOBER 30: Navy Week Special Event Station. The Laurel, Maryland, ARC will operate K3LDE on board USF Constellation, 1200-2200 UTC. Frequencies: primary 7225, 14225 and 21400 per band condx. Request 3 first-class stamps to cover mailing tube and specially designed certificate to Box 259, Anapolis Junction, MD 20701.

NOVEMBER 1983 / $2.50

ham radio • November 1983

SPECIAL ISSUE: RECEIVERS

- Dr. Ulrich Rohde on EMI/RFI receiver design requirements
- optical FM receiver
- time domain reflectometer
- modern communication receiver design
- time and frequency standards
- weekend project: simple shortwave broadcast receiver



BUILD THIS COMPACT SSB RECEIVER

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focus on communications technology

ham radio magazine

NOVEMBER 1983

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contents

REFLECTIONS

Thank You

Thank you for responding to the reader survey published in the September issue. I want you to know how very valuable your comments are to me. You are providing to us at *ham radio* an understanding of your needs, likes, and dislikes.

As we go to press, we have not yet "closed the books" on this survey; survey forms and letters continue to arrive daily in considerable quantity. In reviewing the forms that have been returned, I already see important trends which together with the survey compilation, will be detailed in a future issue. In general, readers are telling us to continue to provide a technically superior magazine and not to succumb to the temptation of trying to offer "something for everybody." This was clearly brought out in one of the many letters that accompanied the survey forms:

I cannot resist writing to tell you how grateful I am that you have gone back to the original aims of ham radio. *I had originally subscribed to the magazine because of the technical content and I am so pleased to see a return to the more technically oriented articles. The fact is that there are already three other magazines devoted to operating news and 'beginner' type construction articles — we don't need one more. I am sure that your general reader response will confirm that there is a need for material that has some body to it and which can be used for future reference....*

Taking all of the responses (so far) into account, I promise to follow a policy I adopted several years ago while editor of another magazine (*rf design*): *always try to inform, not impress.* Now, I have to admit that some of our past articles did not exactly meet that criterion, and I plead guilty with explanation. In the publishing industry it's not unusual for an editor to be working with the content of four or more issues in various stages of development at the same time. But by pushing the schedule ahead with excellent material received from our readers, I'll be able to improve the quality of *ham radio*. I've just finished editing some exciting new material to be included in the early 1984 issues.

While on the subject of articles, I'd like to ask all prospective authors to remember that the most important reason for writing and publishing a manuscript is to communicate — be it an idea or a complete system down to the last diode. Sometimes it helps to ask a friend to read your manuscript with a critical eye. If there's something he or she doesn't understand, chances are that others will have the same problem. Before you start, I firmly believe that it's always a good policy to make a single-sided, single-page outline on the proposed subject. Don't write another word until you really like the outline. It's much easier to modify an outline than to cut and paste or start a manuscript all over after you've gotten into it. By all means, once you like the outline, stick with it, taking one section or thought at a time — in order. This makes the job considerably easier.

The kitchen sink. Yes, I've found a few of them in some of the manuscripts. There is a tendency among some of us (and I am not excluding myself) to try to squeeze everything you possibly can into a five-printed-page article. Please believe me, you can save some material for another article or a book or even an encyclopedia. We at *ham radio* can help out in many ways. Send for our well-written six page *Author's Guide.* It will provide you with many helpful hints for producing your manuscript (hint number 1: type your manuscript, double-spaced). Drop me a line with your outline and I will try to respond ASAP with suggestions.

Just a final word on artwork. Penciled sketches are fine. We normally redraw all schematics, block diagrams, etc., unless you happen to follow our drafting style and produce camera-ready artwork (some authors do). Photographs should be black and white, 35 mm or larger. Use the best photographic techniques you can master (clean backgrounds, good lighting, logical presentation) and don't hesitate to seek professional assistance. Remember, if accepted, your work will adorn the pages of a widely, worldly circulated magazine.

There we have it. I've thrown the kitchen sink into this editorial. Let me just say thanks again for your support. Please keep reading the magazine; we'll keep trying to improve and expand it to meet your needs.

Rich Rosen, K2RR
editor-in-chief

presstop de W9JUV

THE VOLUNTEER EXAM PROGRAM FOR AMATEURS WAS ESTABLISHED officially September 22, when the FCC acted on PR Docket 83-27. The biggest surprise was the Commissioners' decision to use 13 regional Volunteer Examiner Coordinators instead of one national VEC, with a VEC in each U.S. call area plus one each for Alaska, the Pacific islands and the Caribbean. The decision to go with regional VECs is being widely interpreted as a direct slap at the ARRL, whose last-minute introduction of a demand that VECs be compensated for their efforts after assuring the FCC they could handle it gratis caused much consternation at the Commission.

Exam Administration Fees Are Specifically Prohibited by the Report and Order, in a section reportedly written by the FCC's legal staff and based on the enabling legislation. The League is lobbying on Capitol Hill for a bill that would legalize fee collection, but Senator Goldwater has come out strongly against such fees and without his support it's unlikely that it can receive much support in either house.

Three-Person Examining Teams Are Still Required for the Technician and higher class exams. All but Technician will require that all three team members be Extra Class. Specifically, 13 as well as 20 wpm code tests plus Elements 4(A) (Advanced and Extra) and 4(B) (Extra) must be administered by a team consisting of three Extra Class licensees.

Negotiations With Groups Wishing To Become Regional VECs will be opened by the FCC on December 1. Since some areas will respond more quickly and have an acceptable examiner organization in place more rapidly than others, it appears almost certain that the program will be up and running in some parts of the country long before it will be in others.

LAUNCH OF THE STS-9 SPACECRAFT IS STILL SET FOR OCTOBER 28, with W5LFL due to begin his 2-meter operation from space a few days later as outlined in September Presstop. Late-breaking information will be made available via recorded messages on various special phone lines, including ARRL--(203) 666-0688, Westlink--(213) 465-5550, and the Johnson Spaceflight Center--(713) 483-2477. In addition, Electra (Bearcat) is making its toll-free line (800) SCA-NNER available as a mission progress hotline to Amateurs as well as other VHF listeners for the duration of the STS-9 mission.

The Most Up-To-Date STS-9 Information Will Probably Be From W5RRR, the Space Center Radio Club station. It will be on the air before and after working hours plus weekends, using 28600, 21375, 14280, and 3845 kHz, all +/- QRM. Operation on OSCAR 10 is possible.

Retransmission Of Space Shuttle Transmissions By Amateurs has been authorized by the FCC in response to several petitioners. However, Amateurs wishing to perform this service must first get permission from NASA.

FCC ACTION ON THE "NO-CODE" AMATEUR LICENSE is unlikely until early 1984, according to Washington sources. There is also considerable speculation that the rebuff the League took on the Amateur licensing program is a harbinger of a pro No-Code decision when the Commissioners do finally consider that thorny issue.

The Widely Heralded Air Force Letter Opposing No-Code was apparently only a statement by some Air Force MARS people that their MARS appointments (which do include HF band operations) would require CW ability, and was not a statement of official Air Force policy. In addition, no basis has been found for a recently circulated rumor that the CIA had told the FCC that a No-Code license "would not be in the national interest."

2-METER USE BY FISHING BOATS IN PUGET SOUND is concerning Amateurs in the Pacific Northwest. Reminiscent of similar episodes during the height of the CB boom, when some truckers discovered readily available 2-meter rigs offered them a refuge from the bedlam of channel 19, the fishermen are using 2 meters for "private communications channels" to discuss matters inappropriate for the regular marine band or which they want kept secret from others not "in the know." It appears the operation was set up by someone knowledgeable, since the fishermen have pretty well avoided use of active Amateur frequencies.

The FCC Has Been Informed And Is Actively Monitoring the illegal operations. At least a dozen of the pirate stations seem to be active.

When The FCC's Program To Involve Amateurs In Enforcement will get under way seems up in the air at the moment. Despite earlier hopes it would be in operation by this fall, there has apparently been little progress on it in the last few months.

"WB23XYZ" AND "AB84C" WILL BE LEGITIMATE CALLSIGNS for California Amateurs during July and August, 1984. California Amateurs with "6" in their callsigns have been authorized by the FCC to use either "23" (for 23rd Olympiad) or "84" (for 1984) instead of "6" during the period of the Olympic games in California next year.

Amateur Involvement In The Olympics Is Progressing Well, with plans now firm to have Amateur HF stations operating from all three Olympic villages. Tentative agreements are already well along with a number of countries to waive their restrictions on third-party traffic, to enable their Olympic athletes to keep in touch with home via Amateur Radio.

PIZZA ORDERING BY AMATEUR RADIO WAS BRIEFLY LEGALIZED by a recent, short lived FCC policy relaxation. The change, in effect for only a few weeks in September, was another effort to resolve on-going conflicts over what constitutes prohibited communications. It has since been rescinded and the Part 97.114(c) restrictions remain in effect.

comments

cable choice

Dear HR:

The article, "Inexpensive Hardline Connectors," by WB4GCS (May, 1983) was read with considerable interest. Mr. Sanford's connector may work very well in his particular application, but there are many other considerations and problem areas to be evaluated. I will try to outline some of them.

Do not assume all cables are alike. Using 1-inch cable as a reference, there are at least three major varieties with center conductor diameters ranging from 0.19 to 0.24 inch. If scaling to 3/4-inch cable, there are at least five major varieties. Choose the piece that will mate with the center conductor carefully.

The center conductor should be securely captured. I recommend that the inner portion of the barb fitting be threaded. This is similar to what is used on commercial connectors, and will assist in retarding conductor migration.

The center conductor should thread onto the connector by at least 1/2 inch, particularly if any power is contemplated. If one lives in an area of temperature extremes, the use of cable with a solid copper center conductor should be avoided. The different coefficients of expansion between the aluminum sheath and the copper conductor can lead to all kinds of failures. The copper-covered aluminum center conductor was developed to alleviate this. This is a prime reason the commercial connector manufacturers use materials that exhibit expansion coefficients similar to the cable.

Mr. Sanford's paragraph on using an anti-corrosion compound cannot be stressed too strongly. Failure to do so can cause the cable to self-destruct in less than a year in certain environments. (Contact an electrician or electrical supply house for the brand names available in your area.)

Results must be taken with a grain of salt. If a certain homebrew connector works for your particular project, by all means use it. Up to approximately 150 MHz, just about anything will work reasonably well and give a return loss of 14 dB (VSWR 1.5) or better. Commercial connectors readily achieve return losses of better than 25 dB and virtually immeasurable insertion loss. The 1-inch cables that I am familiar with have a loss per 100 feet of from 0.4 to 0.5 dB at 150 MHz and 0.75 to 0.95 dB at 450 MHz. This can go a long way toward putting power where it belongs.

One final caution about the cable. Use only fresh cable, the source of which you are certain. If possible, find out the upper frequency limit of the CATV system in which it is used. Most new systems are operating to 400 MHz or higher and are using cable with excellent characteristics well past 500 MHz. There are, however, many varieties of older design cable that are being passed off to hams by unscrupulous individuals at fleamarkets. Many of those cables deliver horrible performance above about 200 MHz. When going up through the UHF bands, verifiable results require sophisticated test equipment and thorough attention to detail. Even more insidious is the type of cable that does not have the foam dielectric bonded to the sheath and inner conductor. Any water ingress will then migrate completely throughout the cable. This will quickly turn a kW station into QRP level ERP even at 20 meters! For high power, at least 3/4-inch cable should be used to reduce the possibility of high-voltage RF flashover.

I am personally all for the use of 75-ohm hardline and am designing my station for its use. It is produced by the millions of feet, is reliable and reasonably priced even when new. Hams have been getting ripped off for years by sticking to 50-ohm cable. If you don't have the time or mechanical dexterity to produce a connector, contact your local CATV engineer. You will probably be pleasantly surprised to find what is available for the asking.

Carl Huether, KM1H
Pelham, New Hampshire

freebies

Response to our recent offers of supplementary materials has been tremendous — our thanks to all who wrote. Copies of the World Press List, the NASA Tech Brief, and the RTTY-Atari™ program are still available; send a large SASE (with 20¢ in stamps for the press list, 37¢ for the tech brief, and 54¢ for the program) for copies of one or all. A sampling of recent letters follows. — Editor

Dear HR:

I enjoyed the article "RTTY and the Atari™ Computer" by Dave King, K5VUV. Hopefully I can figure out a way to interface into the computer serial port on my Atari 400, rather than use the Atari interface module, which costs more than the computer!

Chuck Hastings, KB3QU
Annapolis, Maryland

Dear HR:

I really enjoyed "RTTY and the Atari™ Computer." Although my recently-purchased computer is a TRS-80C,™ I believe I can use the interface. Please send the program listing.

P.B. Johnson, VE7DHM
Sooke, British Columbia

Dear HR:

Please send me a copy of the NASA tech brief.

Thank you for a fine publication. I am particularly interested in Forrest Gehrke's series on phased verticals, having used and worked with them with moderate success for some time. This is the first definitive article on the subject to appear in the Amateur literature.

Arthur J. Conebeer, W6DRL
Laguna Beach, California

compact SSB receiver

Bigger isn't always better. Here's a small, easy-to-build unit that performs much like a full-sized receiver

This simple, compact receiver has a lot of grown-up features: a built-in speaker, automatic gain control, good sensitivity, wide dynamic range, and the potential for excellent selectivity. Add a small voltage-probe antenna and you can cast off the feedline and take the receiver with you.

Most of the necessary components are readily available and inexpensive; modular design allows you to chose your own packaging. All board layouts, photos, and diagrams are provided, and any builder with modest experience should find construction no problem at all.

circuit description

In many respects, this circuit is similar to others described in recent Amateur literature.[1,2] However, it has some practical features which offer a great deal of flexibility.

Fig. 1 shows the main receiver board. A switchable 20-dB attenuator provides RF gain control to prevent receiver overload. Q1 is a grounded-gate RF amplifier which provides 10 dB of gain ahead of mixer Q2. Q2 is a single-ended MOSFET mixer. This stage is coupled to bandpass filter FL-1 by means of T1, a broadband matching transformer. Either a mechanical or crystal filter can be employed by choosing the appropriate turns-ratio. The filter's output is terminated by resistor R_F. Since the filter is not mounted on the circuit board, physical size is not a factor in filter selection.

Q3 is the receiver's gain IF stage. Gain is controlled by a simple audio-derived AGC system. Diodes replace Q3's source resistor in order to bias gate-2 negative with respect to gate-1. This extends the AGC attenuation range.[3] IF transformer T2 is capacitively coupled to product detector Q4. The transformer's secondary is not used. A toroidal LC circuit can replace this transformer for non-standard IF frequencies. IF frequencies from 455 kHz to 9 MHz and beyond can be employed without board modification.

Product detector Q4 is an active circuit which provides audio pre-amplification ahead of the gain control. U1, the audio amplifier, is an LM-386. This IC provides a voltage gain of 200 and delivers 400 mW of power into an 8-ohm load. A 10-ohm series resistor in the output line protects miniature 200 mW speakers from damage.

The AGC system is a simple audio-derived limiter.[4] A diode samples the output of U1 and sends a negative voltage to dc amplifier Q5/Q6. The output of Q6 is set for a resting bias of +4 volts under no-signal conditions. When a strong signal appears, this voltage drops to as low as +0.5 volts, reducing Q3's gain. Resistor R_D sets the AGC sampling level. The value of R_D is selected for best AGC action. Capacitor C_A is optional, but recommended for SSB operation since it slows release time.

The VFO shown in **fig. 2** is a near-copy of a W7ZOI/W5IRK circuit.[5] This design provides excellent performance. The Hartley JFET oscillator drives a single MOSFET buffer/amplifier. VFO-output is coupled to the mixer through a broadband trans-

By Rick Littlefield, K1BQT, Box 114, Barrington, New Hampshire 03825

The completed receiver can be packaged to suit the builder. This 20-meter version features a signal-strength meter and a built-in voltage probe antenna system. The illuminated frequency pointer is an LED that has been filed to shape and polished.

former. Builders desiring a thorough treatment of the design along with temperature compensation information should refer to current editions of the *ARRL Handbook*.

Figs. 3A, **B**, and **C** show BFO circuits. These provide plenty of output at the high input impedance presented by Q4. The 455 kHz version employs a ceramic resonator instead of a crystal. These devices are considerably cheaper, and much easier to "rubber." Frequency adjustment is accomplished with a small 60 pF trimmer. The high-frequency version uses standard crystals, and oscillates easily in the 3-12 MHz range.

Fig. 4 shows an optional tuned voltage probe antenna circuit. The telescoping rod antenna is coupled directly to the Hi-Z end of L1. A dual-gate MOSFET provides pre-amplification and impedance matching for the rod. Pre-amp output is transformed to 50 ohms through a broadband transformer. The longer the rod antenna, the more broadband the response. On 20 meters, a 2-foot (60-cm) rod allows coverage of the entire phone band. On 75, a 4-foot (120-cm) rod covers around 50 kHz. The antenna and pre-amp can be mounted in a receiver case, or remoted from the exterior of a structure or vehicle.

Finally, **fig.** 5 shows a simple regulator and pilot

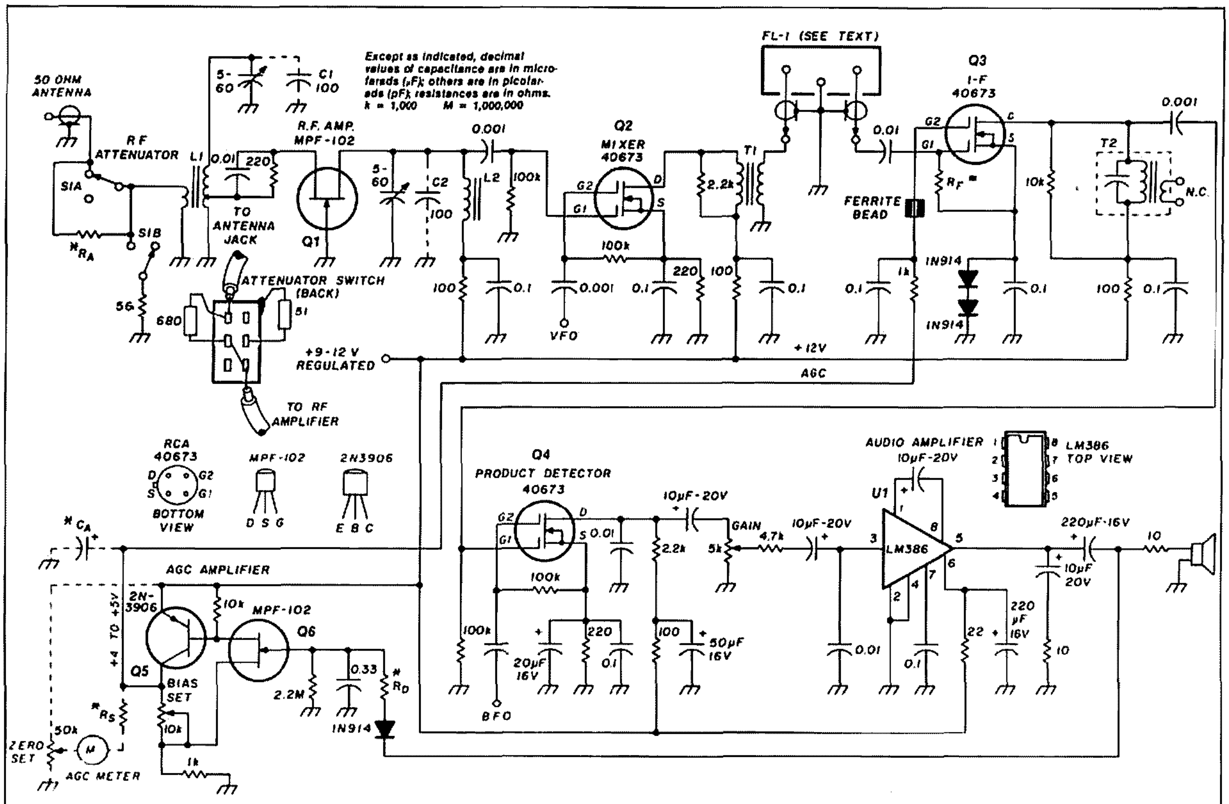
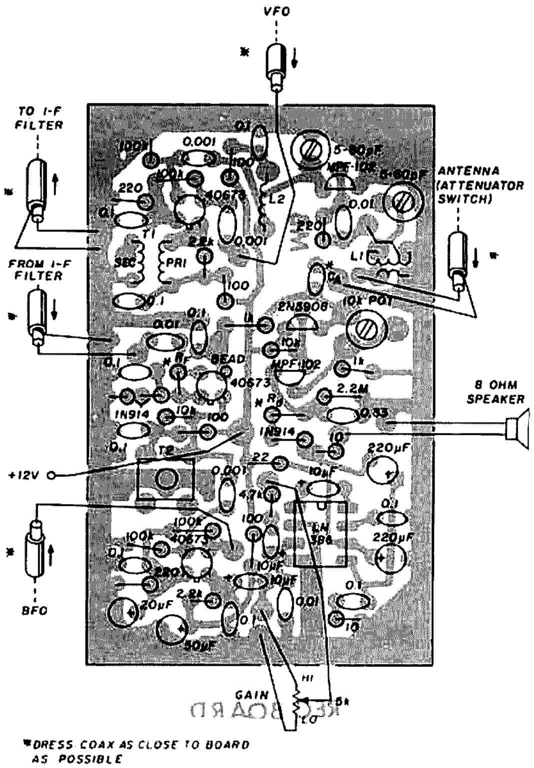


fig.1A. Main receiver board schematic.



80-meter operation	
C1,C2	100 pF
L1	2 feet of No. 36 on a T37-2 core: tap 1/3 up from grounded end, add 5-turn link for antenna on ground end
L2	same as L1 except omit tap and link
20-meter operation	
C1,C2	not used
L1	28 turns of No. 28 on T37-2 core: tap at 8 turns, add 2-turn link for antenna on grounded end
L2	same as L1 except omit tap and link
455-kHz Mechanical filter i-f plan	
FL1	455-kHz mechanical filter. Collins F455FD series, or Kokusai MF-455-ZL-21H (MHZ Electronics). Install recommended resonator capacitors at the filter
RF	2.2K for most filters
T1	1:1 turns ratio broadband transformer: lightly twist together two 1-1/2-foot lengths of No. 36, and wind on a FT37-43 core
T2	J.W. Miller No. 2067, or virtually any 455 kHz miniature transformer from the junk-box (secondary not used)
9 MHz crystal filter i-f plan	
FL1	KVG XF-9B, Atlas 9.0 USB/CW (MHZ Electronics), McCoy, etc. Provide resistive termination per manufacturer's instructions. Terminate input at filter, terminate output by means of resistor "rf" on gate No. 1 of the i-f amplifier stage
RF	same value as filter termination resistor, usually between 300 and 820 ohms
T1	2:1 turns ratio broadband transformer: primary, 15 turns No. 28 on FT37-43 core; secondary, 8 turns No. 28
T2	any miniature 10.7 MHz. i-f transformer (secondary not used). Will need additional 15-22 pF across primary to resonate at 9 MHz. Use pads provided on underside of the board
other variable component values	
Ca	Sets AGC speed. Not needed for fast action, 2.2-4.7 μF for slower action
Ra	Sets level of rf attenuation; 680 ohms provides approximately 20 dB of attenuation
Rd	Sets AGC drive. May vary from 10-50K, see text for set-up information
Rs	Sets meter sensitivity. 50 μA movement requires 20K; less sensitive movements require less

fig. 1B. Parts layout for main receiver board.

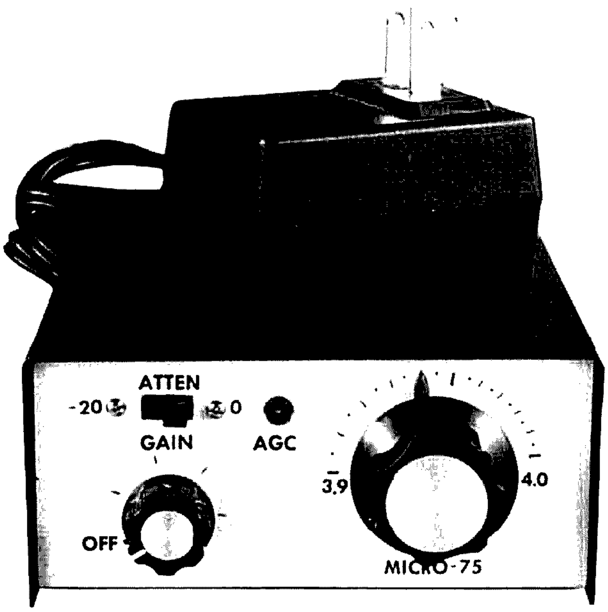
LED circuit. The LM-7812 holds the operating voltage at 12 volts, protects the modules from damage, and keeps noise out of the system.

construction

The entire receiver chain, RF amplifier through audio, is contained on one main circuit board. Because oscillators require shielding, they are built separately. Optional circuits such as the regulator and antenna pre-amp are also separate, since some builders may choose to omit them.

Original artwork for my boards was prepared on transparent acetate stock using Radio Shack rub-ons. Boards were prepared with the General Cement positive developer system and pre-sensitized board. You can use this same system by applying lift-film to pull the board patterns from this article. Pre-etched boards are also available from Radiokit.[6]

Component density is fairly high on all of these boards. Miniature parts should be used wherever possible to prevent crowding. Choosing small tantalum audio coupling capacitors, low voltage bypasses, 1/4- or 1/8-watt resistors, and compact elec-



Good things can come in small packages. This 75-meter net monitor is almost dwarfed by its AC adapter.

trolytics will result in attractive and uncluttered boards. Shielded and unshielded wires should also be small in diameter and flexible. Most components are available at Radio Shack stores or by mail-order through Radiokit.[6] I reduced construction costs considerably by drawing on a parts inventory built from junked circuit boards and surplus grab-bags.

Coils are much more difficult to prune after they are installed on the circuit board. For that reason, all tuned circuits are wound, tacked together with solder, and checked for resonance with a grid-dip meter prior to actual board construction. A length of hook-up wire is used to link-couple the grid-dipper to the toroids. Each tuned circuit is then marked and set aside for later installation.

Oscillators are constructed first, since they are needed to test the main board. Special attention is given to mechanical stability, especially while constructing the VFO. Cement firmly in place anything that can move or vibrate. Upon completion, check for oscillation and proper output level.

When selecting a VFO main-tuning capacitor, look for a "ball-bearing" type with a good vernier drive (either built-in or added on). Variables and vernier drives are available from several sources including Radiokit and BCD Electro.[7] Select fixed capacitors for the VFO tank with the main tuning capacitor mounted in the receiver case. This provides mechanical stability during component substitution. Values for C1, C2, and C3 are juggled until the desired tuning range and dial linearity is obtained. During this process, a frequency counter and pocket calculator are very helpful. The counter provides an accurate measure of the oscillator frequency, and the calculator adds and subtracts the IF frequency to give the actual receive frequency. If a counter is not available, a good general coverage receiver will suffice.

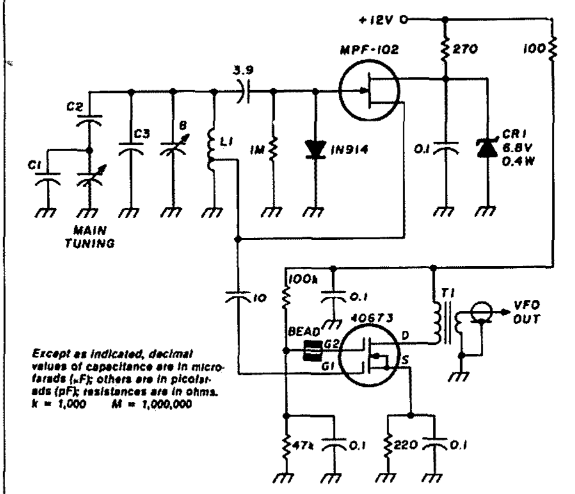


fig. 2A. VFO board schematic.

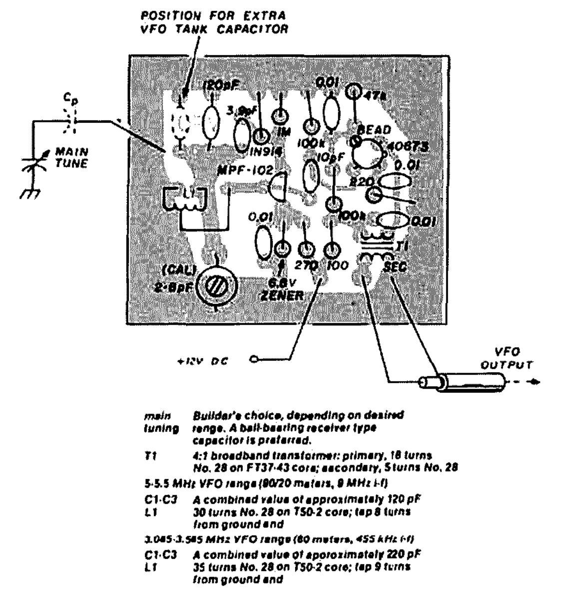


fig. 2B. VFO board and parts layout.

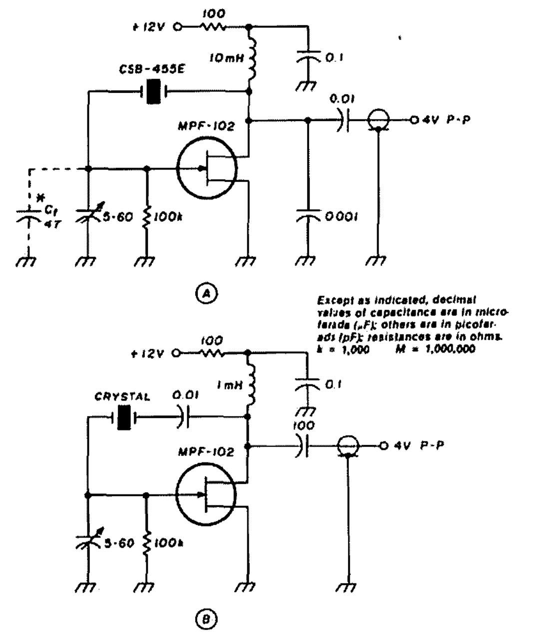
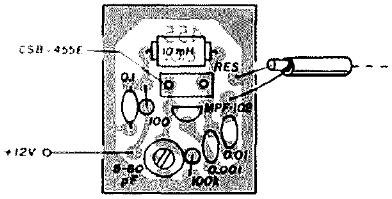


fig. 3A. 455 kHz BFO board schematics.

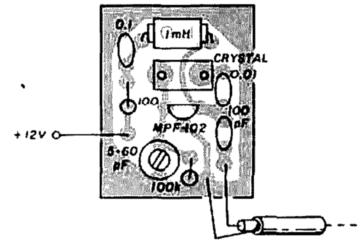
After final installation of all frequency determining components, mount the VFO board in place and mark dial calibrations on the front panel. While full shielding of the VFO is desirable, this may be difficult in compact packages like the "Micro-75" shown in the photo on page 13. In practice, some VFO leakage does not appear to degrade performance.

Construction of either BFO board is simple. After assembly and testing, a "can" or small box is constructed to shield the board. Unlike the VFO, the BFO must be fully shielded to prevent birdies and common-mode detector noise. Tin flashing, two-sided board, or aluminum all make good boxes. Two small holes are needed for the leads, plus an access hole for tuning the trimmer. Install the shielded BFO in the receiver case.



C1 — 47 pF, may be required to pull oscillator to 453.6 kHz. Add to foil side of the board
RFC — 10 mH miniature choke, J.W. Millar 70F102A1 or equivalent
crystal — Radio Shack 455 kHz ceramic resonator (272-1303)

fig. 3B. Ceramic resonator 455 kHz BFO.

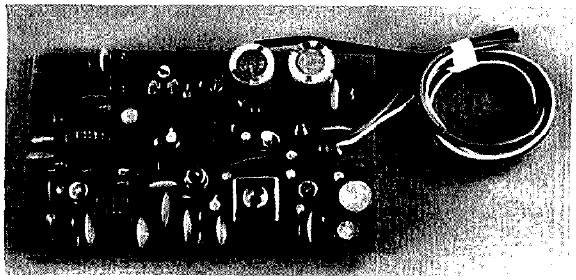


RFC — 1 mH miniature choke, J.W. Millar 70F103A1 or equivalent
crystal — 30 pF load capacitance, miniature holder

fig. 3C. Crystal oscillator BFO.

Since the main board is more complex than the others, it is constructed one stage at a time. Start with the audio and AGC sections, and work back to the RF amplifier. This keeps the process orderly, and allows stage-by-stage inspection. It also leaves installation of vulnerable toroid inductors until last.

A number of frequency plans are possible for this receiver. Here are some construction tips for the two versions I have built.



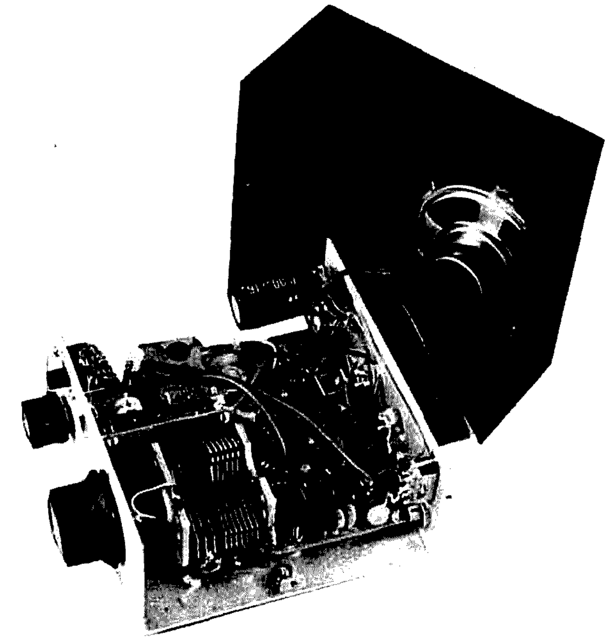
With the exception of the oscillators, all receiver circuitry is contained on the main board. Interconnecting leads for the speaker, volume control, and DC power are salvaged rainbow-wire.

The "Micro-75" receiver is based on a 455 kHz IF. Most 455 kHz designs use a mechanical filter. A 1:1 input transformer and a 2.2 K value for R_F matches most of these devices. If external resonating capacitors are required, install them at the filter. A data chart for most Collins filters appears in current editions of the *ARRL Handbook*. T2 can be any miniature 455 kHz can. Avoid using the 455 kHz IF above 40 meters, since insufficient image rejection will be available at the higher frequencies.

The 20-meter portable uses a 9 MHz IF. High-frequency IF's generally employ a crystal filter. This requires a 4:1 input transformer and a 300 to 800 ohm output termination. When termination numbers are not available, use a 560 ohm resistor for R_F. For transformer T2, use any 10.7 MHz miniature can and add 15 - 22 pF of padding to lower the resonant frequency. This capacitor can be installed beneath the board on the extra set of pads provided for this purpose.

Coil data is supplied for 80- and 20-meter operation. For operation on 40, 30, and 15 meters, a survey of other receiver articles will provide LC values close enough to get started. The T37-2 forms are quite small. Preparing the 20-meter inductors is easy enough, but concentration and a steady hand are needed to wind the 75-meter versions. I used No. 36 wire for these because it was available, but there is room on the form to substitute No. 34. All toroid coils and transformers are glued to the board after installation to prevent excess movement and lead breakage. Note that the 100 pF padding capacitors are installed for the 75-meter front-end only.

Oscillator inputs and mounting pads for resistor R_D require solder-pins on the top side of the board. (Small flea-clips or discarded resistor leads are fine for this purpose.) After all components are mounted, wires for interconnections are installed. Cut all leads



Total shielding of the VFO may not be possible in small packages like this one. This does not seem to degrade performance. The BFO should always be fully shielded. Note the bandpass filter mounted on the rear panel to save interior space.

on the long side to facilitate dressing during final assembly. Ground-loops are always a possibility when modules are interconnected. To prevent this, ground shielding at one end only.

receiver check-out and alignment

Final assembly can begin after the main board is completed. In my prototypes, the bandpass filter is externally mounted on the back panel of the receiver case. This saves interior space and eliminates the need to fabricate a mounting bracket. All jacks, the RF attenuator, gain control, and speaker are mounted prior to board installation. The board itself is mounted on short spacers with No. 2-56 hardware. Leads are then trimmed to length and connected to their destinations.

Initial testing and alignment is quite simple. Connect a resistance substitution box or a calibrated 50K pot to the terminals provided for resistor "Rd". A value of 10K is fine for initial testing. Advance the receiver gain control to about 3/4 volume, and apply power to all three modules. A very soft hiss from the speaker indicates that no serious shorts are present. Adjust the "bias-set" for a resting AGC voltage of +4 volts, as measured at the top end of the potentiometer. Tune the IF through its range, looking for a slight noise peak to indicate resonance. Finally, peak the RF amplifier. If everything is working and the band is open, the receiver will come to life. RF trimmers should show two signal-peaks as they are rotated through 360 degrees. This confirms that resonance is within the trimmer range, and not off to one

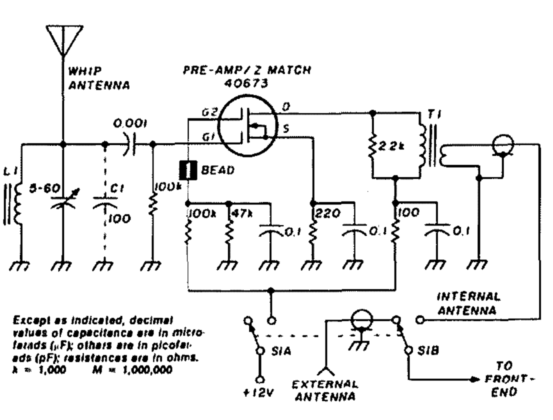
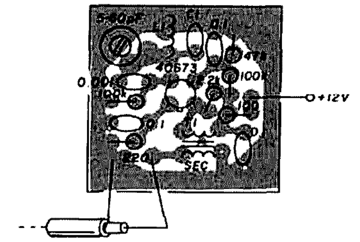
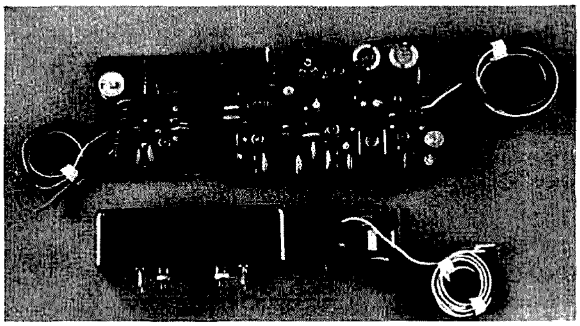


fig. 4A. Voltage probe antenna board schematic.



C1	100 pF on 80 meters, not used on 20 meters
L1	Same as corresponding L2 on main receiver board
T1	6:1 turns ratio broadband transformer; primary, 18 turns No. 28 on FT37-43 core; secondary, 3 turns

fig. 4B. Parts layout for voltage probe antenna.



The main board, two oscillator boards, and a bandpass filter are the basic modules required to build one of these receivers. Interconnecting wires are salvaged rainbow-wire and mini-audio cable.

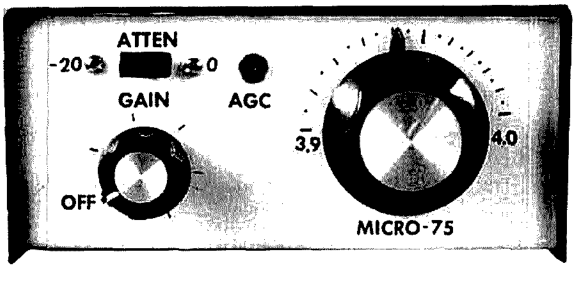
side. Stagger tune the RF stage to provide even sensitivity across the entire tuning range of the radio.

To determine the correct value for R_D, tune in an extremely strong signal and vary the resistance until AGC action is smooth. Too much AGC produces overshoot. This is a condition where the AGC over-responds, producing a "pumping" effect. Too little AGC allows the audio amplifier to go into distortion. The best value should fall somwhere between 10K and 50K. Install the nearest standard-value resistor.

This design has one quirk that might cause alarm during the testing phase. Extremely strong signals will sound fuzzy at low volume, yet miraculously clear up when the gain is increased. This is because AGC is derived from the audio output, and the AGC's ability to control the IF is pre-empted by the manual gain control at very soft listening levels. Fuzziness at low volume indicates overloading. A 20-dB RF attenuator is included in the receiver-chain to correct this condition. In practice, it becomes second nature to switch in the attenuator when the band is open and signals are strong.

options

The most important addition is the regulator circuit described earlier in fig. 5. This provides cheap insurance for a project well done, and cleans up dirty power sources like automotive electrical systems and inexpensive AC adapters. Since these are the sources I use most often, both of my receivers are regulated. Mount the LM-7812 on the interior of the back panel, and use the positive output lead as a tie point for all of the module power leads. If hum persists with your AC adapter, it could be that the ripple is dipping below the regulating range of the LM-7812. If this happens, install a 10-ohm resistor in series with adapter's plus-lead or get a new adapter with more output voltage.



Simplicity of design is reflected in the front panel of this 75-meter net monitor. In place of a signal-strength meter, an LED peak-indicator illuminates at full audio output to indicate AGC action.

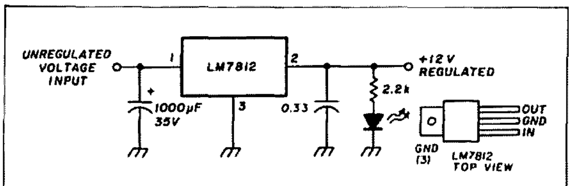


fig. 5. Regulator schematic.

The optional meter circuit shown in **fig. 1** is not a full blown S-meter, but does measure relative signal strength. Almost any sensitive movement can be adapted to this circuit with the appropriate value of R_S. Use care while experimenting, since accidently grounding the AGC line destroys the 2N3906.

The voltage probe antenna circuit is a great addition when the receiver is going to be taken along as a portable. This circuit board is mounted inside the cabinet of my 20-meter prototype and connected to a short collapsible whip that extends through the top of the case (the 75-meter receiver uses one as an external accessory). The pre-amp components are very similar to those used in the receiver front-end, and the same techniques apply for construction. A DPDT switch on the back panel applies power and brings the pre-amp on line for portable use. The antenna trimmer is accessible through the back of the cabinet, since peaking is quite critical and may require re-adjusting from time to time. (It's a thrill to hear VK's and ZL's rolling in on a 2-foot [60 cm] whip while sipping coffee at the kitchen table!)

conclusion

This is a very functional receiver design, easy to construct from available parts. The unusually flexible circuit allows selection of alternate frequency-plans without board modification. Many of the features one would expect to find in a full-sized communications receiver are included.

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3. G. Woodward, W1RN, editor, *The Radio Amateur's Handbook*, 60th edition, ARRL, Newington, Connecticut, pages 8-33 to 8-37.
4. Rick Littlefield, K1BQT, "Construct an Audio Amplifier with AGC for Your Simple Receiver," *QST*, April, 1983.
5. G. Woodward, W1RN, editor, *The Radio Amateur's Handbook*, 60th edition, ARRL, Newington, Connecticut, pages 8-40 to 8-42.
6. *Etched circuit boards*, parts, and kit for the "Compact SSB Receiver" are available from Radiokit, Box 411, Greenville, New Hapmshire 03048.
7. Variable capacitors and other parts available from BCD Electro, P.O. Box 119, Richardson, Texas 75080.

ham radio

designing a modern receiver

Achieve high performance through careful selection of LO and IF frequencies using IMD charts

Most general-coverage receivers have, up till now, been designed around commercially available IF filters. The popular 9 MHz approach performs well but exhibits a variety of internally generated spurious products when used in a general coverage mode. The chart in **table 1** shows some of these products and how they impact on received frequencies. (Products produced in the premixing schemes of local oscillators have not been considered in this example.)

While the design trend has been to use up-conversions with first IF's higher than the highest frequency to be received (typically 35 percent higher for improved image rejection), performing a system design for the frequency scheme of a multi-conversion receiver has been considered a complicated mathematical analysis beyond the ability of the average Amateur. This need not be so; this article provides the reader with the tools necessary for understanding the design process more fully.

In a receiver, mixers provide undesired output products in addition to their sum and difference frequencies. These products are called intermodulation products. This phenomenon is complicated by the increased front-end bandwidth requirement referred to as general coverage as well as by the IF bandwidth requirement. If a multi-mixer situation exists, such as in a multiconversion receiver, the problem is further aggravated, as initial unwanted products from one mixer combine with those of another, creating a multitude of "birdies" (unwanted interference that sounds like the whistling of a bird) at the final IF output. Regardless of whether a receiver is dedicated or general coverage, the problem of intermodulation products has to be carefully understood and weighed against system parameters so that the fewest possible "birdies" are internally generated and heard within the passband of the receiver.

predicting IMDs

Let's look at some analytical tools the system designer uses to determine these products. Assume we're going to design a fixed-frequency receiver for 70 MHz (**fig. 1**). With a local oscillator of 90 MHz, the receiver will have a first IF of 160 MHz using an up-conversion mixing technique. (The second conversion of this receiver is not discussed here in order to simplify this case.) To use the mixer product chart (**fig. 2**) normalized frequencies must be calculated. Dividing the design frequency by the local oscillator frequency generates the first normalized number:

$$\frac{f_1}{f_2} = \frac{70}{90} = 0.778$$

Dividing the first IF by the local oscillator frequency generates the second normalized number:

$$\frac{f_{i\text{-}f}}{f_2} = \frac{160}{90} = 1.778$$

With this information and the mixer product chart, find the locus point (the intersection of a system of lines which satisfies one or more given conditions) for the two ratios, as shown in **fig. 3**. The chart in **fig. 2** shows all products produced not only by the fundamentals, but also by multiples of the signal and oscillator frequencies present in the mixer stage, and correspond to the second, third, fourth, fifth, and sixth harmonics of the two mixed signals.

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The **order** of the **product** is determined by the sum of the harmonic orders involved. For example, $5f_1 \pm 2f_2$ is a seventh-order product (regardless of the mathematical operation involved) because it involves the fifth harmonic of f_1 combined with the second harmonic of f_2. Higher-order products are also present, but they are usually of a sufficiently low level so as not to cause problems. Any line that crosses the locus point corresponds to a product which is identified on the edge of the chart.

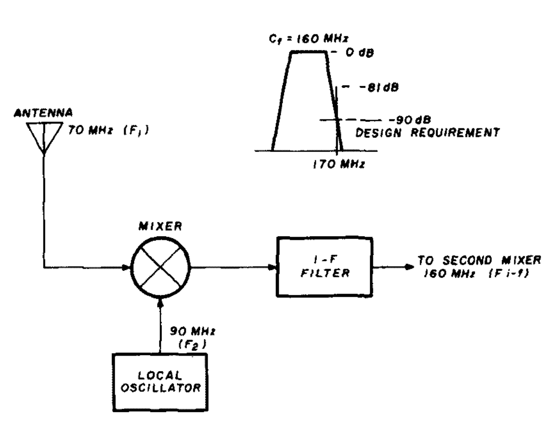

fig. 1. Finding in-band intermodulation products dictates choice of IF center frequency as well as the shape factor of the IF filter, in order to meet system requirements.

Values of f_1 and f_2 can be substituted and the interference can be anticipated and avoided. If the locus point is examined closely, one can be assured that there are no in-band products in this case, but analyzing the areas adjacent to the locus point indicates some out-of-band spurs (spurious, unwanted products) which will have to be suppressed by the IF filter to the level specified in the requirement. By knowing their order (given by the chart), their predicted amplitude can be found (in our case, 170 MHz). The seventh and ninth-order products ($5f_1 - 2f_2$ and $5f_2 - 4f_1$) are predicted to be 81 dB below the IF level (typical manufacturer prediction). The IF filter will have to provide 9 additional dB of attenuation at 170 MHz to accomplish a system requirement of 90 dB as shown in **fig. 1**. A simpler method of finding these products can be achieved by using the charts in **tables 2** and **3**. The chart shown in **table 2** is for mixers used in an additive mode (B + A) where A and B are the mixing frequencies and $B > A$. The chart shown in **table 3** is for mixers used in a subtractive mode (B − A) with the same conditions applying. If using the same example, and substituting f_1 for A and f_2 for B, the same ratio can be obtained.

$$\frac{f_1}{f_2} = \frac{A}{B} = \frac{70}{90} = 0.778$$

We then use **table 2**, since the mixer in our example operates in the additive mode, and find the corresponding products as indicated in **fig. 4** ($5A$-$2B$ and $5B$-$4A$). If the numerical values of A and B are inserted in these formulas, the same resultant values can

table 1. Examples of spurious frequencies present in receiver using 9 MHz IF.

OPERATING FREQUENCY MHz	LOCAL OSCILLATOR FREQUENCY MHz	BFO FREQUENCY MHz	LOWEST ORDER SPURIOUS PRODUCT EQUAL TO i-f	TYPICAL CARRIER LEVEL (μV) FOR 10 dB $\frac{S+N}{N}$		
				ADJACENT FREQUENCIES MAXIMUM	ADJACENT FREQUENCIES TYPICAL	SPURIOUS FREQUENCY TYPICAL
3.0025	12.0025	9.0025	3LO-3BFO	0.3	0.2	2.0
3.6030	12.6030	9.0025	5LO-6BFO	0.3	0.2	2.0
5.4035	14.4035	9.0025	5LO-7BFO	0.3	0.2	6.0
6.0033	15.0033	9.0025	3LO-4BFO	0.3	0.2	30.0
9.0025	18.0025	9.0025	LO-BFO	0.4	0.2	Receiver Blocked
11.2550	20.2550	9.0025	4LO-8BFO	0.4	0.2	8.0
12.0050	21.0050	9.0025	3LO-6BFO	0.4	0.2	300 for 3dB $\frac{S+N}{N}$
13.5000	4.5000	8.9975	2LO	0.5	0.3	Receiver Blocked
14.99916	5.99916	8.9975	3LO-BFO	0.5	0.3	8.0
18.0000	9.0000	8.9975	LO	0.5	0.3	Receiver Blocked
20.9975	11.9975	8.9975	3LO-3BFO	0.6	0.4	0.5
21.5970	12.5970	8.9975	5LO-6BFO	0.6	0.4	2.0
23.3965	14.3965	8.9975	5LO-7BFO	0.6	0.4	8.0
23.99667	14.99667	8.9975	3LO-4BFO	0.6	0.4	2.0
25.1960	16.1960	8.9975	5LO-8BFO	0.6	0.4	4.0
26.9975	17.9975	8.9975	LO-BFO	0.6	0.4	30.0
29.2450	20.2450	8.9975	4LO-8BFO	0.6	0.4	6.0

be obtained with this method, which is usually preferred. If this receiver were not designed for a fixed frequency, you can imagine what a job it would be to evaluate all the higher-order products generated by using this method.

computers speed the analysis

Today, computer programs are used successfully to help designers anticipate potential problems. In our example, a TI-59 programmable calculator is used to perform this tedious task. (A program listing is included in **table 4** for those wishing to work out their own problems.)

This program finds all combinations of (m × LO) ± (n × RF) and prints those frequencies that fall in the center of the IF by actually indicating "IF" in the printout. Those frequencies that fall within the predetermined IF bandwidth but are not exactly in the center are also reported in the printout by the indication "**BW**" (**fig. 6**). The sample program in **fig. 6** shows how our 70 MHz fixed-frequency receiver may be analyzed using this method. If the 90 MHz LO is entered into the user-defined key A, the 70 MHz RF into B, the 160 MHz IF into C, and the IF bandwidth we wish to analyze into D (50 MHz), a report is obtained by depressing key E′, indicating that we are ready to run the program with the entered data as shown in **fig. 5**. If a mistake occurred in the process of entering the information, new data can be entered by repeating the above process with no alteration to the actual program. We can now run our analysis by depressing key E. A complete list of products will be automatically printed as shown in **fig. 6**. The process takes approximately four minutes to analyze all cases

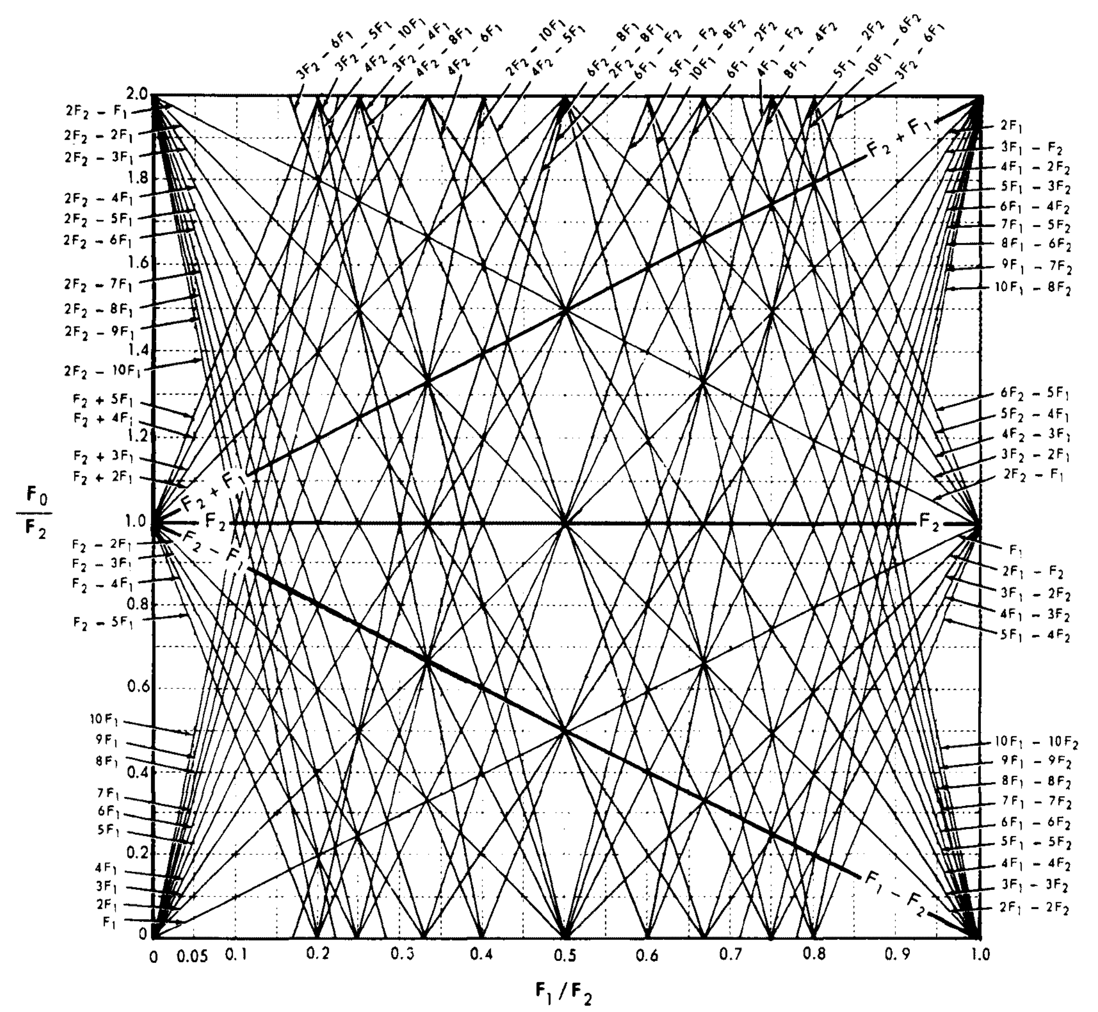

fig. 2. Intermodulation products chart.

of m and n within the 50 MHz bandwidth. This may seem to be a long time, but not if we compare the time expended in manually searching the product chart in **fig. 2**.

This program can be recorded on two magnetic cards. For those with a TI-59, **table 4** shows the actual listing of the intermodulation products program. Partitioning (OP 17) is 479.59. **Table 5** lists the procedures necessary for running the program. The amplitude of the undesired products identified depends on their particular order number (m + n). Most products of the seventh order or higher will be at least 60 dB down from the IF level, and are usually not considered to cause problems. Unless different instructions are entered, the program will automatically calculate all products to the twelfth order ($6 \times LO$) $\pm$ ($6 \times RF$) (no user *inputs to* A′ and B′ are required). If a different resolution is desired, the two keys should be addressed accordingly. The execution time of this program is a direct function of the product order and the IF bandwidth required by the user.

a system design for a general coverage communications receiver

In the following pages we will consider a system design for a general coverage HF receiver with a wide input bandwidth (28 MHz). Unlike the dedicated single frequency receiver analyzed in the above, this wideband receiver presents a considerably more

table 2. Mixer IMD chart, additive mode.

I-F FREQUENCY OUTPUT = B+A (CONDITIONS B>A)

$R = \frac{A}{B}$	ORDER 1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
0.000	B	B − A	B ± 2A	B ± 3A	B ± 4A	B ± 5A	B ± 6A	B ± 7A	B ± 8A	B ± 9A	B ± 10A	B ± 11A	B ± 12A	B ± 13A	B ± 14A
0.072															2B-13A 15A
0.077														2B-12A 14A	
0.083													2B-11A 13A		
0.091												2B-10A 12A			
0.100											2B-9A 11A				
0.111										2B-8A 10A					
0.125									2B-7A 9A						
0.143								2B-6A 8A							
0.154															3B-12A 14A-B
0.167							2B-5A 7A							3B-11A 13A-B	
0.182													3B-10A 12A-B		
0.200						2B-4A 6A						3B-10A 11A-B			
0.222											3B-8A 10A-B				
0.250					2B-3A 5A					3B-7A 9A-B					4B-11A 13A-2B
0.273														4B-10A 12A-2B	
0.286									3B-6A 8A-B						
0.300													4B-9A 11A-2B		
0.333				2B-2A 4A				3B-5A 7A-B				4B-8A 10A-2B			
0.364															5B-10A 12A-3B
0.375											4B-7A 9A-2B				
0.400							3B-4A 6A-B							5B-9A 11A-3B	
0.429										4B-6A 8A-2B					
0.445													5B-8A 10A-3B		
0.500			2B-A 3A			3B-3A 5A-B			4B-5A 7A-2B			5B-7A 9A-3B			6B-9A 11A-4B
0.555														6B-8A 10A-4B	
0.571											5B-6A 8A-3B				
0.600								4B-4A 6A-2B							
0.625													6B-7A 9A-4B		
0.667					3B-2A 4A-B					5B-5A 7A-3B					7B-8A 10A-5B
0.715												6B-6A 8A-4B			
0.750							4B-3A 5A-2B							7B-7A 9A-5B	
0.800									5B-4A 6A-3B						
0.833											6B-5A 7A-4B				
0.858													7B-6A 8A-5B		
0.875															8B-7A 9A-6B
1.000		2B 2A		3B-A 3A-B		4B-2A 4A-2B		5B-3A 5A-3B		6B-4A 6A-4B		7B-5A 7A-5B		8B-6A 8A-6B	

table 3. Mixer IMD chart, subtractive mode.

I-F FREQUENCY = B-A (CONDITION B>A)

$R = \frac{A}{B}$	ORDER 1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
0.000	B	B ± A	B ± 2A	B ± 3A	B ± 4A	B ± 5A	B ± 6A	B ± 7A	B ± 8A	B ± 9A	B ± 10A	B ± 11A	B ± 12A	B ± 13A	B ± 14A
0.063															15A
0.067														14A	
0.072													13A		
0.077												12A			
0.083											11A				2B-13A
0.091										10A				2B-12A	
0.100									9A				2B-11A		
0.111								8A				2B-10A			
0.125							7A				2B-9A				
0.133															14A-B
0.143						6A				2B-8A				13A-B	
0.154													12A-B		
0.167					5A				2B-7A			11A-B			
0.182											10A-B				3B-12A
0.200				4A				2B-6A		9A-B				3B-11A	
0.214															13A-2B
0.222									8A-B				3B-10A		
0.231														12A-2B	
0.250			3A				2B-5A	7A-B				3B-9A	11A-2B		
0.273												10A-2B			
0.286							6A-B				3B-8A				
0.300											9A-2B				4B-11A
0.308															12A-3B
0.333		2A				2B-4A 5A-B				3B-7A 8A-2B				4B-10A 11A-3B	
0.364													10A-3B		
0.375									7A-2B				4B-9A		
0.400					4A-B				3B-6A			9A-3B			
0.416															11A-4B
0.428								6A-2B				4B-8A			
0.445											8A-3B				5B-10A
0.455														10A-4B	
0.500	A			3A-B	2B-3A		5A-2B	3B-5A		7A-3B	4B-7A		9A-4B	5B-8A	
0.545															10A-9B
0.555												8A-4B			
0.571									6A-3B				5B-8A		
0.600						4A-2B				4B-6A				9A-5B	
0.625											7A-4B				6B-9A
0.667			2A-B				3B-4A	5A-3B				5B-7A	8A-5B		
0.700															9A-6B
0.715										6A-4B				6B-8A	
0.750					3A-2B				4B-5A			7A-5B			
0.778														8A-6B	
0.800							4A-3B				5B-6A				
0.833									5A-4B				6B-7A		
0.858											6A-5B				7B-6A
0.875													7A-6B		
1.000				2B-2A		3B-3A		4B-4A		5B-5A		6B-6A		7B-7A	

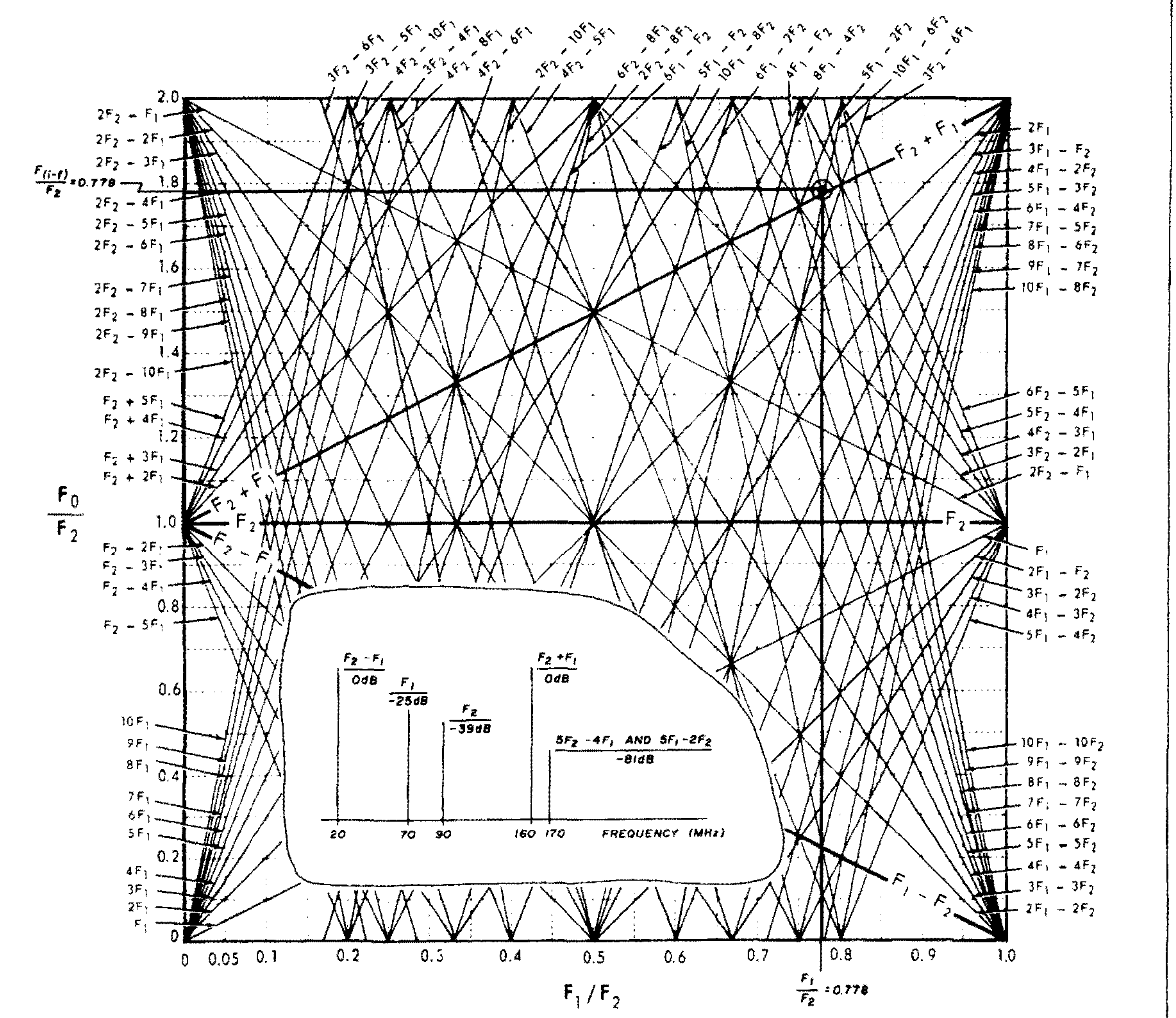

fig. 3. The intersection of the vertical and horizontal lines corresponding to $\frac{f_1}{f_2}$ and $\frac{f_{(IF)}}{f_2}$ (locus point of normalized frequencies) indicates the in-band intermodulation products. The out-of-band spurious outputs that happen to be in the vicinity of the IF frequency can also be verified by looking at the products adjacent to the locus point.

complex product analysis problem, because of the many different cases that could be created within the input bandwidth. The problem is further increased if a double-conversion approach is used, since products generated in the first IF can multiply in the second IF. The designer should use good judgment in the initial choice of frequencies, since no computer or chart can take the place of good engineering procedures.

The design objectives are a communication receiver covering 2 to 30 MHz, with good image rejection, having a minimum of unwanted products. Looking at **fig. 7**, a double conversion approach is considered with an up-conversion first IF compatible with commercially-available monolithic crystal filters at 75 MHz. A phase-locked synthesizer used as the local

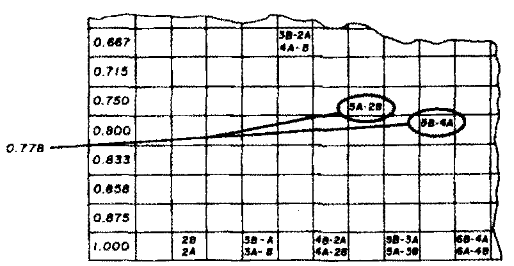

fig. 4. The same results can be obtained as with our previous example by using the intermodulation chart from table 2. Ratio $\frac{A}{B} = 0.778$ which points to a 7th and 9th order product (5A-2B and 5B-4A).

oscillator for the first mixer must be tunable in 10 kHz steps over this range. Fine tuning is achieved in the second conversion stages with another synthesizer which provides frequency resolution of 100 Hz within the 10 kHz steps.

A 9 MHz second IF was chosen because of the availability of good crystal-lattice filters at that frequency. Two sideband filters are used in this IF. The system will be analyzed using the previously described charts; the TI-59 computer program will be used to perform the calculations.

Fig. 8 shows the mathematical model for this receiver. An RF signal within a 2.000 to 30.000 MHz range is injected into the first mixer, where it subtracts from the first local oscillator frequency and

CONDITIONS	
6.	× LO
6.	× RF
90.	LO
70.	RF
160.	IF
50.	BW

fig. 5. Depressing Key E′ will print out all conditions entered by the operator.

table 5. User instructions for the TI-59 intermodulation products program.

STEP	DESCRIPTION	ENTER	PRESS
1	Input the highest LO harmonic of interest*	m	A′
2	Input the highest rf harmonic or interest*	n	B′
3	INPUT LO FREQUENCY	LO	A
4	INPUT rf FREQUENCY	rf	B
5	INPUT i-f FREQUENCY	i-f	C
6	INPUT BANDWIDTH Centered at i-f	BW	D
7	PRINT CONDITIONS ENTERED		E′
8	RUN PROGRAM		E

Progam finds all combinations of mxLO ± n × rf and frequencies that are equal to tha i-f chosen (=i-f) and those that lie within the i-f bandwidth (*BW).

*If no input, a sixth order will be automatically analyzed.

USER DEFINED KEYS

A	LO
B	rf
C	i-f
D	BW
E	RUN PROGRAM
A′	m
B′	n
C′	NOT USED
D′	NOT USED
E′	CONDITIONS REPORT

table 4. Listing of the TI-59 program for intermodulation products. Partitioning is 479.59.

000	76	LBL	049	43	RCL	098	01	1	147	80	80	196	92	RTN	245	06	06
001	11	A	050	03	03	099	94	+/-	148	43	RCL	197	00	0	246	43	RCL
002	42	STO	051	85	+	100	44	SUM	149	19	19	198	00	0	247	17	17
003	03	03	052	43	RCL	101	02	02	150	69	OP	199	00	0	248	69	OP
004	91	R/S	053	02	02	102	97	DSZ	151	04	04	200	76	LBL	249	04	04
005	76	LBL	054	65	×	103	09	09	152	43	RCL	201	10	E′	250	43	RCL
006	12	B	055	43	RCL	104	19	D′	153	05	05	202	98	ADV	251	06	06
007	42	STO	056	04	04	105	43	RCL	154	69	OP	203	98	ADV	252	69	OP
008	04	04	057	95	=	106	12	12	155	06	06	204	43	RCL	253	06	06
009	91	R/S	058	50	I×I	107	42	STO	156	98	ADV	205	23	23	254	43	RCL
010	76	LBL	059	42	STO	108	02	02	157	92	RTN	206	69	OP	255	20	20
011	13	C	060	05	05	109	85	+	158	00	0	207	03	03	256	69	OP
012	42	STO	061	71	SBR	110	01	1	159	00	0	208	43	RCL	257	04	04
013	06	06	062	01	01	111	95	=	160	71	SBR	209	24	24	258	43	RCL
014	91	R/S	063	15	15	112	42	STO	161	01	01	210	69	OP	259	07	07
015	76	LBL	064	43	RCL	113	09	09	162	80	80	211	04	04	260	69	OP
016	14	D	065	01	01	114	91	R/S	163	43	RCL	212	69	OP	261	06	06
017	42	STO	066	65	×	115	43	RCL	164	18	18	213	05	05	262	98	ADV
018	07	07	067	43	RCL	116	06	06	165	69	OP	214	43	RCL	263	98	ADV
019	91	R/S	068	03	03	117	32	X:T	166	04	04	215	15	15	264	91	R/S
020	76	LBL	069	75	-	118	43	RCL	167	43	RCL	216	69	OP	265	00	0
021	15	E	070	43	RCL	119	05	05	168	05	05	217	04	04	266	00	0
022	43	RCL	071	02	02	120	67	EQ	169	69	OP	218	43	RCL	267	00	0
023	06	06	072	65	×	121	01	01	170	06	06	219	01	01	268	00	0
024	75	-	073	43	RCL	122	45	45	171	98	ADV	220	69	OP	269	00	0
025	43	RCL	074	04	04	123	43	RCL	172	92	RTN	221	06	06	270	76	LBL
026	07	07	075	95	=	124	13	13	173	00	0	222	43	RCL	271	16	A′
027	55	÷	076	50	I×I	125	32	X:T	174	00	0	223	16	16	272	42	STO
028	02	2	077	42	STO	126	43	RCL	175	00	0	224	69	OP	273	01	01
029	95	=	078	05	05	127	05	05	176	00	0	225	04	04	274	42	STO
030	42	STO	079	71	SBR	128	77	GE	177	00	0	226	43	RCL	275	11	11
031	13	13	080	01	01	129	01	01	178	00	0	227	02	02	276	85	+
032	85	+	081	15	15	130	32	32	179	00	0	228	69	OP	277	01	1
033	43	RCL	082	01	1	131	92	RTN	180	43	RCL	229	06	06	278	95	=
034	07	07	083	94	+/-	132	43	RCL	181	15	15	230	43	RCL	279	42	STO
035	95	=	084	44	SUM	133	14	14	182	69	OP	231	21	21	280	08	08
036	42	STO	085	01	01	134	32	X:T	183	04	04	232	69	OP	281	91	R/S
037	14	14	086	97	DSZ	135	43	RCL	184	43	RCL	233	04	04	282	76	LBL
038	68	NOP	087	08	08	136	05	05	185	01	01	234	43	RCL	283	17	B′
039	68	NOP	088	19	D′	137	22	INV	186	69	OP	235	03	03	284	42	STO
040	68	NOP	089	43	RCL	138	77	GE	187	06	06	236	69	OP	285	02	02
041	68	NOP	090	11	11	139	01	01	188	43	RCL	237	06	06	286	42	STO
042	68	NOP	091	42	STO	140	60	60	189	16	16	238	43	RCL	287	12	12
043	68	NOP	092	01	01	141	92	RTN	190	69	OP	239	22	22	288	85	+
044	76	LBL	093	85	+	142	00	0	191	04	04	240	69	OP	289	01	1
045	19	D′	094	01	1	143	00	0	192	43	RCL	241	04	04	290	95	=
046	43	RCL	095	95	=	144	00	0	193	02	02	242	43	RCL	291	42	STO
047	01	01	096	42	STO	145	71	SBR	194	69	OP	243	04	04	292	09	09
048	65	×	097	08	08	146	01	01	195	06	06	244	69	OP	293	91	R/S

CONDITIONS		CONDITIONS	
3.	× LO	0.	× LO
6.	× RF	2.	× RF
150.	* BW	140.	* BW
2.	× LO	1.	× LO
5.	× RF	1.	× RF
170.	* BW	160.	= IF
5.	× LO	2.	× LO
4.	× RF	0.	× RF
170.	* BW	180.	* BW
4.	× LO		
3.	× RF		
150.	* BW		

fig. 6. Depressing Key E on the TI-59 will run the intermodulation product program within the specified conditions indicating the same problem areas 5RF-2LO and 5LO-4RF as previously determined with the intermodulation charts. Other problems are also indicated within the IF bandpass but are not considered in our example for reasons of simplicity.

produces a 75.000 MHz IF. This local oscillator is configured as a synthesizer operating from 77.000 to 105.000 MHz in 0.01 MHz (10 kHz) steps. This dictates the bandwidth of the first IF to be 10 kHz minimum, from 74.995 to 75.005 MHz in order for the second local oscillator to be able to provide fine tuning in the second IF. A 75 MHz tandem monolithic filter from Piezo-Technology Inc. can be used in this application. If $RF = A$ and $LO = B$, two ratios $\frac{A}{B}$ can be created: R minimum and R maximum.

$$R_{MIN} = \frac{2.000}{77.000} - 0.025$$

$$R_{MAX} = \frac{30.000}{105.00} = 0.285$$

We will use the mixing product chart (**table 3**) for subtraction since our IF is 75 MHz, and find the entire band between R_{MIN} and R_{MAX} as shown in **fig. 8**. Any product indicated within this band could be a potential problem for the corresponding received frequency.

A look at the chart indicates a series of problems (7A, 6A, 5A, 4A, 3A, 2B-5A, with the worst one at 3A). If the TI-59 program is used, we can verify this case as shown in **fig. 8**. At first we can say that the third harmonic of one of the two mixing signals could be quite powerful and could indeed produce a problem, but a closer look at the system indicates that the offending frequency *A*, is actually a received frequency and chances are very good that a distant 25 MHz station has a level of insignificant third harmonic (75 MHz) appearing at the antenna of our receiver.

The problem is further diminished by our receiver's preselector, which greatly attenuates at 75 MHz. The same conditions apply to the other products indicated by the chart. They present even a better case since they are further removed from the received frequencies. This is a case where judgment is more important than all our tools, which are only used to warn of possible problems.

The case would be different, however, if the B signal were the offender, as 3B would have been the third harmonic of the local oscillator, which can be of relatively high amplitude and cause interference.

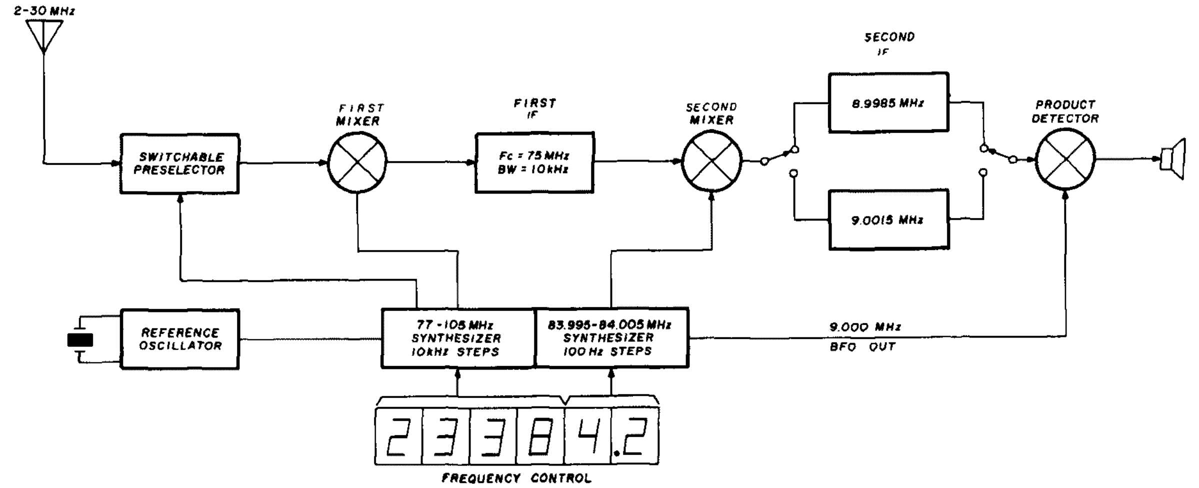

fig. 7. Communications receiver block diagram. The first IF is 75 MHz with a second IF of 9 MHz. A two-loop synthesizer is used to provide coarse and fine tuning.

The filtered first IF range of 74.995 to 75.005 MHz is further mixed with the second local oscillator operating in 0.0001 MHz (100 Hz) steps over the 10 kHz range of 83.995 to 84.005 MHz, providing fine tuning for the receiver. If R_{MIN} and R_{MAX} are found for the second IF a new band of interest can be located on our chart, as shown in **fig. 8**.

$$R_{MIN} = \frac{74.995}{83.995} = 0.89285076$$

$$R_{MAX} = \frac{75.005}{84.005} = 0.89286352$$

Since the range to be covered is only 10 kHz, the band is very narrow and in reality is expressed by the same number (0.892) because the chart extends to only three decimal places. The IF is centered at 9.000 MHz and its bandwidth is determined by the two single-sideband filters. For simplicity, the minimum corner frequency (−3 dB) of the lower filter and the upper corner frequency for the higher filter were chosen, determining a total bandwidth (−3 dB) of 0.0051 MHz (5.1 kHz for both sidebands). Fig. 8 clearly indicates that there is no problem except for a thirteenth order product which can be ignored. Since the band is so narrow and close to the 1.000 ratio which presents quite a few problems, the computer

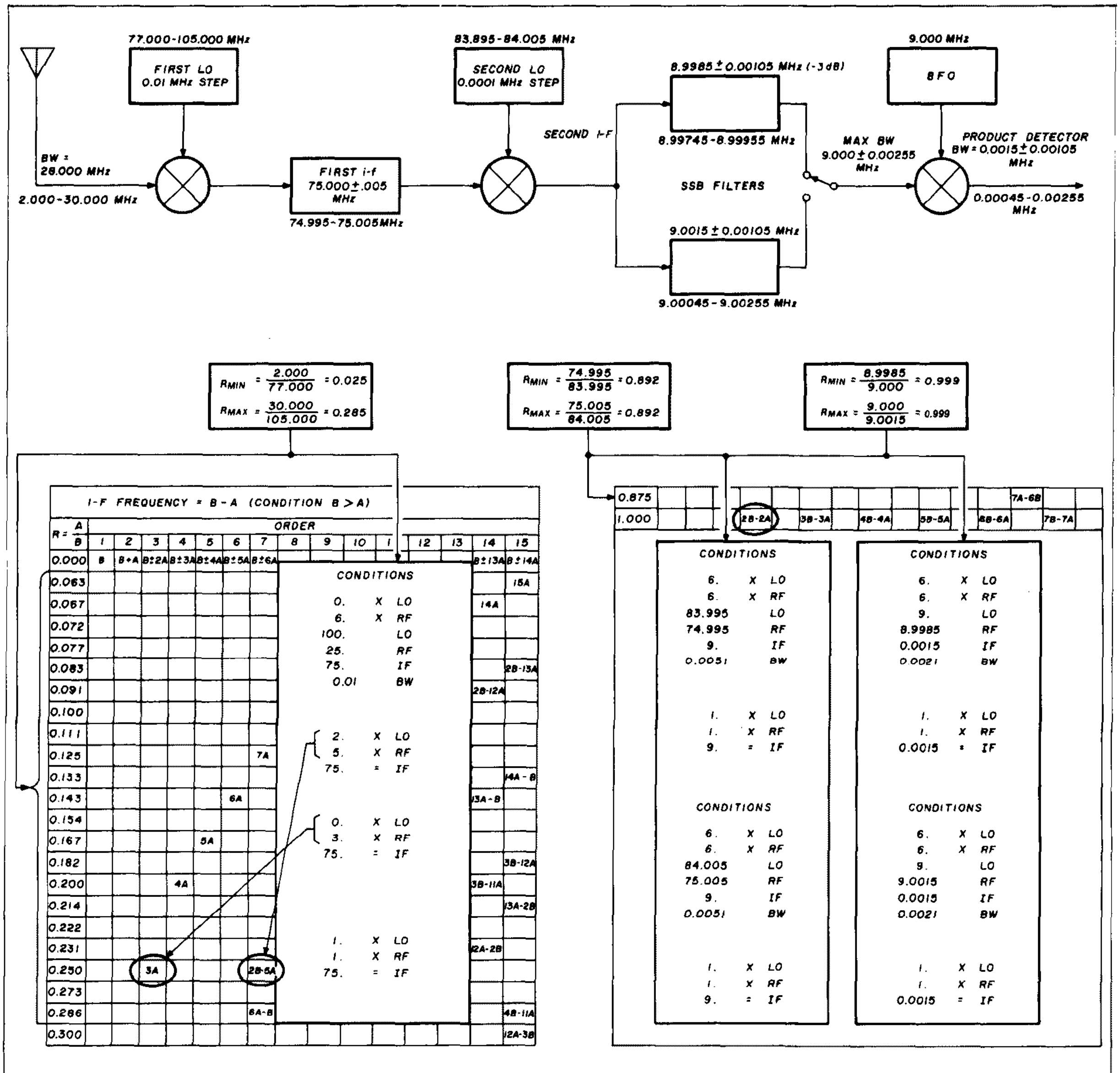

fig. 8. System analysis for a double conversion, general-coverage communications receiver with a first IF at 75 MHz and a second IF at 9 MHz.

CONDITIONS	
7.	× LO
7.	× RF
83.995	LO
74.995	RF
9.	IF
0.0374	BW
1.	× LO
1.	× RF
9.	= IF

fig. 9. TI-59 program indicates no intermodulation problem in the second IF of the double-conversion communications receiver. The −60 dB bandwidth (0.0374 MHz) of the filters was used for a worst-case analysis.

is used to completely insure safety. Unless othewise instructed, the program will not point out the thirteenth product as it is programmed to only calculate products to the twelfth order. If the order is increased to 14, the computer will indicate that there is no case for a 7A-6B within the ratio range.

In analyzing the second IF in fig. 8, a total bandwidth of 0.0374 MHz (37.4 kHz) is considered to insure complete freedom from intermodulation products within the slopes of the 8-pole filters used in this IF (37.4 kHz is the −60 dB total bandwidth of the two filters), as shown in **fig. 9**. It can be seen from this analysis that the charts can be used only as a guideline. For a more in-depth analysis, the TI-59 program or some other means of calculation must be used to obtain precise answers.

Up to this stage no real interference problems have been encountered. The last conversion is from 9.000 MHz ± 0.00255 MHz to audio between 0.00045 to 0.00255 MHz (450 Hz to 2.550 kHz) in both single-sideband filter cases. The conversion takes place in a third mixer (product detector), as shown in the example (fig. **10**).

$$R_{MIN} = \frac{8.9985}{9.000} = 0.999833$$

$$R_{MAX} = \frac{9.000}{9.0015} = 0.999833$$

These identical ratios locate the intermodulation band to be analyzed as very close to the **1**.000 ratio in the chart and with a −3 dB bandwidth of 0.0021 MHz (2.1 kHz). No problems are found for either one of the single-sideband filters. However, if the −60 dB bandwidth (0.0374 MHz) of the **8**-pole filters is used for the TI-59 computer program, the 2B-2A problem circled in fig. **8** becomes evident at the 3 kHz point in the slopes of our filters. This is true for either sideband filter as shown in fig. **11** and the resulting audio distortion can be cured only by improving the shape factor of the filters, in our case by doubling the number of poles to 16 for each one of the single-sideband filters.

Another way to cure this problem would be to introduce a lowpass filter in the audio portion of our receiver which will cut off all frequencies beyond the 2.55 kHz which is the highest frequency passed by the filters. A practical cut-off point would be at 2.8 kHz. The first method is preferred, however, because it also provides better adjacent-channel rejection, improving overall receiver selectivity.

conclusion

In performing this analysis for the design of a double-conversion general-coverage communica-

CONDITIONS		CONDITIONS	
6.	× LO	6.	× LO
6.	× RF	6.	× RF
9.	LO	9.	LO
8.9985	RF	9.0015	RF
0.0015	IF	0.0015	IF
0.0038	BW	0.0038	BW
2.	× LO	2.	× LO
2.	× RF	2.	× RF
0.003	* BW	0.003	* BW
1.	× LO	1.	× LO
1.	× RF	1.	× RF
0.0015	= IF	0.0015	= IF

fig. 10. The TI-59 program indicates a 2B-2A product within the bandwidth of the 8-pole single sideband filters.

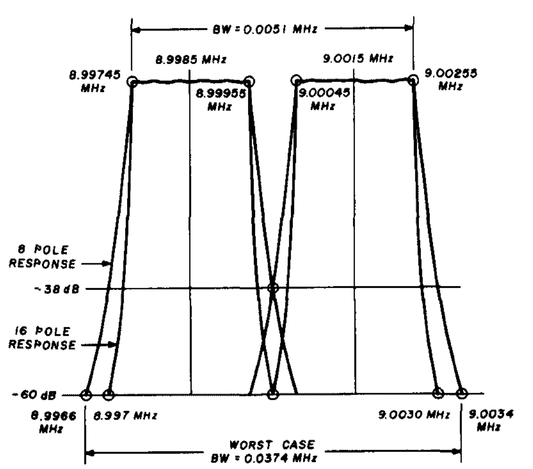

fig. 11. Bandwidth characteristics of the 8-pole single-sideband filters used in the communications receiver. 16 pole filters could be used to eliminate a 2B-2A audio product as well as to improve the total selectivity of the receiver.

tions receiver, we have shown that the design meets the requirements which introduces a minimum of intermodulation problems. We can now proceed with confidence to the circuit design of our receiver. All frequency mixing techniques used in communications receivers generate unwanted products within their outputs, and the fact remains that any configuration chosen is simply the best compromise in the opinion of the designer. The problem of intermodulation distortion within a receiver's scheme can be minimized by performing a careful analysis, such as explained in this article.

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ham radio

time and frequency standards: part 1

Accurate measurement requires accepted standards, precision electronics, and an understanding of atomic physics

Radio Amateurs are probably more concerned about frequency than about any other electrical parameter. However, when discussing frequency, it is important to remember that the reciprocal of frequency is *period*, and that this is expressed in units of time. Because of this relationship, accurate frequency standards must be based on accurate time standards. This two-part article gives a brief historical overview of this field, examines the more common methods of determining frequency and time standards, discusses the advantages and limitations of each, and also succinctly describes some commercial equipment employed in these areas. Part 1 addresses VLF single-frequency comparison techniques; part 2 will cover multiple-frequency VLF techniques and other more advanced systems.

The World Administrative Radio Council (WARC) has set aside five bands for standard frequency and the time signal emissions as shown in **fig. 1**. The carrier frequencies for such broadcasts should be maintained accurately enough so that the average daily fractional frequency deviations from these internationally designated standards for measurement of time intervals do not exceed $\pm 1 \times 10^{-10}$. The map in **fig. 2** shows numerous radio stations used for time-frequency determination (TFD). Stations that provide strong signal transmissions to North America are listed in **table 1**. Other stations that can be received elsewhere in the world are listed in table 2.

LF time and frequency systems

The majority of American and international "VLF" standards stations transmit within the VLF (10 to 30 kHz) and LF (30 to 300 kHz) bands, respectively.

In the early part of this century, both systems were used for long-range communication between colonies and parent countries, and by navies for general trans-oceanic communication. Then, as now, the advantages of VLF and LF systems included reliability with very little signal attenuation, even over vast distances.

In many instances these systems were replaced with HF systems that used smaller antennas. But many novel VLF antenna systems have been built; some consist of long cables strung across valleys or volcanic craters, from towers several hundred meters tall. One such system in Cutler, Maine, radiates 1 MW of power using a "top hat" supported by twenty-six masts, each approximately 300 meters in height. This installation covers over two square kilometers and has a radial ground system of buried copper wire that totals over three million meters in length.

Interest in LF band communications was revived

Band No.	Designation	Frequency Range
4	VLF (Very Low Frequency).	20.0 kHz ± 50 Hz.
6	MF (Medium Frequency).	2.5 MHz ± 5 kHz.
7	HF (High Frequency).	5.0 MHz ± 5 kHz. 10.0 MHz ± 5 kHz. 15.0 MHz ± 10 kHz. 20.0 MHz ± 10 kHz. 25.0 MHz ± 10 kHz.
9	UHF (Ultra High Frequency).	400.1 MHz ± 25 kHz. (satellite).
10	SHF (Super High Frequency).	4.202 GHz ± 2 MHz (satellite space to earth). 6.427 GHz ± 2 MHz (satellite earth to space).

fig. 1. International standard time and frequency radio assignments.

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during World War II, with the evolution of the Radux navigational system, in which low-frequency carriers demonstrated exceptional stability.

Since the mid-1950s there has been great progress in the development of LF communications. Work by Pierce, Mitchell, Essen, and Crombie showed a 100 to 1000 time improvement in frequency compared to existing HF techniques. Early in the 1960s, more stable atomic frequency standards began to take the place of crystal oscillators, and confidence in the

table 1. "Standard" VLF stations receivable in the United States (10.0-30.0 kHz).

station	frequency	location
GBR	16.0 kHz	Rugby, England
NAA	17.8 kHz	Cutler, Maine
NPG	18.6 kHz	Jim Creek, Washington
NPM	23.4 kHz	Laulaulei, Hawaii
WWVL	20.0 kHz and 19.9 kHz	Fort Collins, Colorado
NSS	21.4 kHz	Annapolis, Maryland
NBA	24.0 kHz	Balboa, Canal Zone

table 2. Worldwide VLF stations.

station	frequency	location
JG2AS	40.0 kHz	Kemingawa, Japan
OMA	50.0 kHz	Podebzady, Czechoslovakia
WWVB	60.0 kHz	Fort Collins, Colorado
MSF	60.0 kHz	Rugby, England
HBG	75.0 kHz	Pranginis, Switzerland
DCF-77	77.5 kHz	Mainflingen, Federal Republic of Germany
CYZ-40	80.0 kHz	Ottawa, Canada
FTA-91	91.15 kHz	St. Andre de Corcy, France

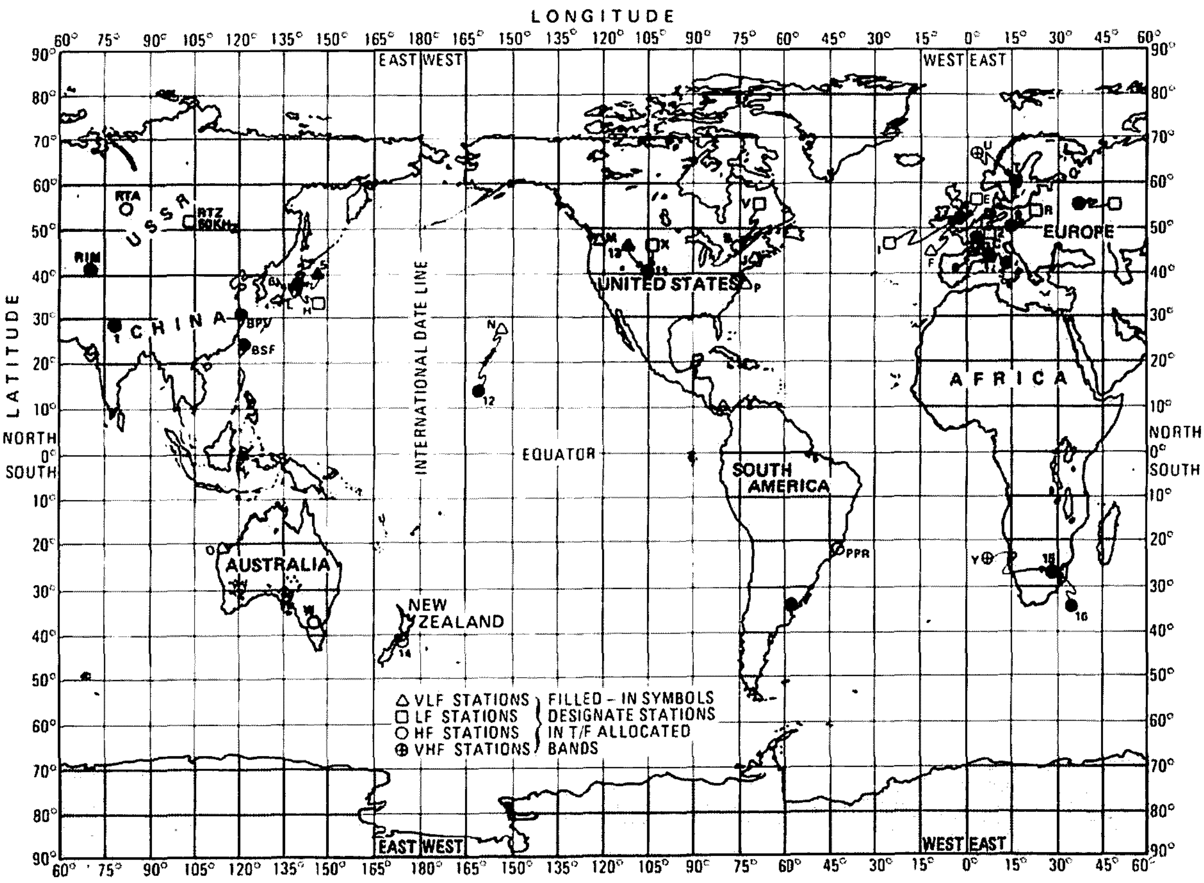

fig. 2. Worldwide location of broadcasting stations useful for TFD.

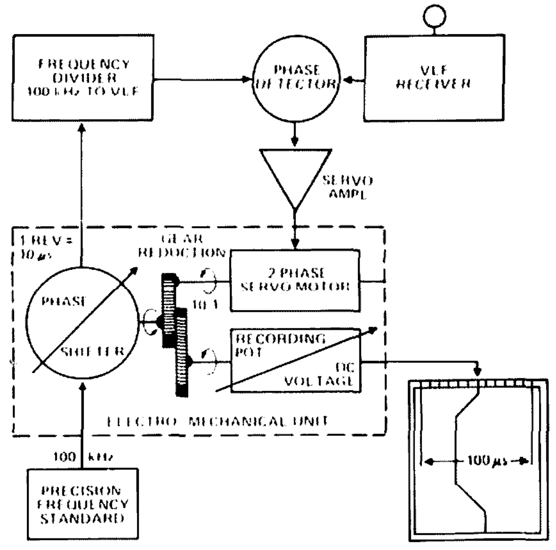

fig. 3. A typical electromechanical VLF single-frequency comparator.

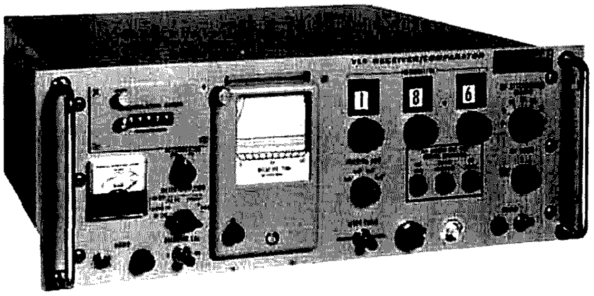

fig. 4. The Fluke Model 207/205 VLF receiver/comparator.

new system was soon rewarded when Pierce, Winkler, and Corke demonstrated that a transatlantic phase comparison could be made on a 16-kHz carrier to within 2 microseconds. Today, most FTDs are made by atomic frequency standards and referenced to a coordinated international time base. This has seen the realization of economical and reliable dissemination of frequency to within several parts per 10^{11} in a 24-hour period.

VLF single frequency comparison

An older, electromechanical VLF method is shown in **fig. 3**. It uses the principle of a servo-driven phase shifter continuously phase locking a synthesized signal from the local standard to the received VLF signal. A linear potentiometer's output, connected to a constant voltage source, generates a voltage related to the phase shifter's position. The recorder shows the amount of phase shift experienced by the local synthesized signal and whether or not it agrees with the phase of the received signal. An early instrument that used this principle was the Fluke model 207/205 VLF receiver/comparator (see **fig. 4**). A more modern piece of equipment designed for the same purpose is the Tracor model 900A VLF/LF receiver shown in **fig. 5**, which operates in the VLF band from 20 to 25 kHz, and compares the phase of a local frequency standard with the received carrier of a frequency-stabilized transmitter. With this instrument, a local standard can be checked to within ten parts per billion.

A related instrument manufactured by Tracor is the model 527E frequency difference meter, shown in **fig. 6**. This solid-state instrument measures frequency differences instantly and has a built-in meter to provide signal-quality assessment regardless of whether the two signals (the reference and signal frequencies) are the same. Since this device is used in the calibration and determination of time and frequency, it is assumed that the two measured frequencies are relatively close to one another. The 527E has a scale that determines signal difference magnitude in the parts per 10^7 to 10^{11} range. An external recorder connector on the back of the instrument is available so that the internally generated dc voltage that is produced in proportion to frequency difference can be recorded. Consequently, this instrument can be used to adjust two oscillators to the same frequency, measure frequency differences between two oscillators, offset one oscillator from another by a given amount, and determine the short and long-term drift of an oscillator.

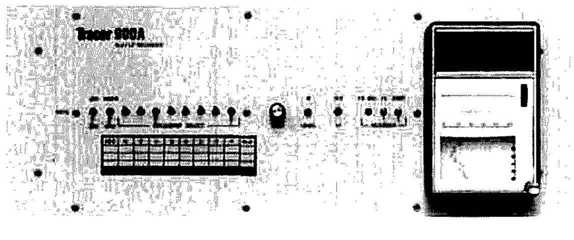

fig. 5. The Tracor VLF/LF receiver (Model 900A).

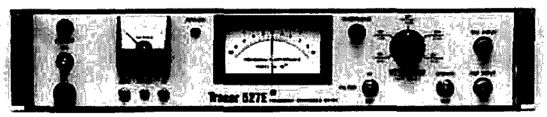

fig. 6. The Tracor frequency difference meter (Model 527E).

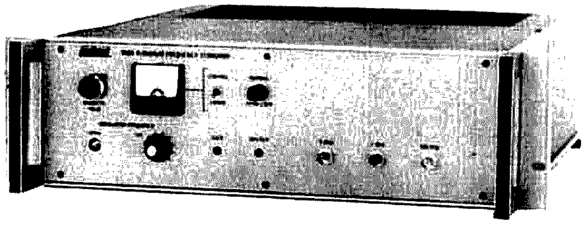

fig. 7. The Tracor rubidium frequency standard, Model 308-A.

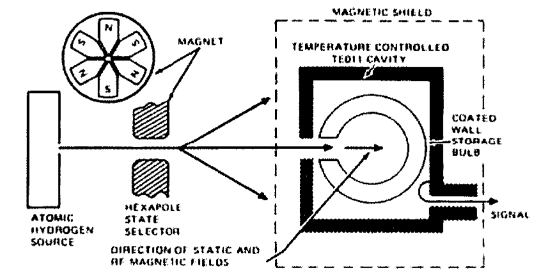

fig. 8. Diagram of the hydrogen maser.

Many labs in which extremely accurate and drift-free frequency standards are required use the Tracor 308-A to determine rubidium 87 frequency standards. This instrument, shown in fig. 7, is set at the factory for 10 parts per billion accuracy and has a month-long drift stability term of better than 30 parts per billion. Used in astronomy, navigation, metrology and communications, this instrument has an MTBF (mean time between failures) of better than 25,000 hours; compare this to many airborne military electronic pieces of equipment with MTBFs of less than 500 hours.

With the advent of solid-state electronics and mass production techniques these seemingly exotic pieces of equipment are not unduly expensive, considering the cost of comparable units ten years ago. The VLF/LF receiver with antenna is priced at approximately $2500; the frequency difference meter, $4800; and the rubidium 87 frequency standard, $14,500.

Why is the rubidium 87 standard so much more expensive than the others? First, let's take a look at atomic frequency standards in general.

atomic frequency standards

Atomic frequency standards employ an atomic resonance device with a frequency source such as a voltage-controlled oscillator phase-locked to it. These devices fall into two categories, active and passive. Cesium beam tubes and rubidium vapor gas cells are passive devices, whereas hydrogen masers are active devices, or resonators.

The hydrogen maser. This — the most stable of all known sources — provides a frequency that is well defined without any need for comparison with an external standard, as we have done previously. (For a quick visual explanation of the operation of the hydrogen maser, see **fig. 8.**) Unfortunately, this piece of equipment is large, expensive, power consuming, and at best suited only to laboratory use at this time. Its use is consequently limited to specialized applications requiring extreme accuracy as well as extraordinary stability. The hydrogen maser uses a beam of hydrogen atoms directed through a highly nonhomogeneous magnetic field that selects atoms in states of higher energy and allows them to proceed into a quartz bulb. Here, a tuned microwave cavity allows atoms to make random transits; the atoms reflected at each encounter within the bulb walls. While undergoing many collisions with the walls, the atoms' effective interaction times with the microwave field is lengthened to about 1 second. During this interaction process, the atoms tend to "relax" and release energy to the microwave field within the tuned cavity. This field also stimulates more atoms to radiate, and a steady-state maser reaction is achieved.

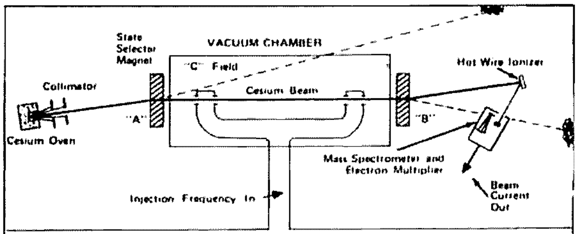

fig. 9. Schematic of the cesium beam tube.

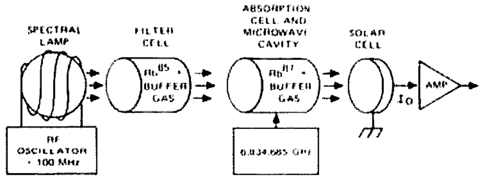

fig. 10. Rubidium vapor frequency standard assembly block diagram.

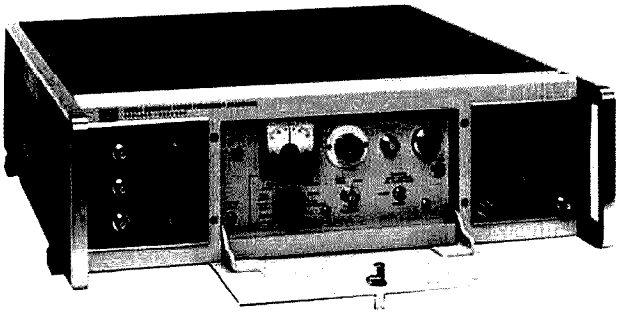

fig. 11. The Hewlett-Packard rubidium vapor frequency standard, Model 5065A.

Cesium beam standards. The basis for operation of the cesium beam standard is the Cesium 133 atom. This system yields an accurate frequency of 9,192,631,700.0000 Hz and is relatively impervious to external electric and magnetic field disturbances. (Refer to **fig. 9** for a visual explanation of the cesium beam standards.) As the Cesium 133 atoms leave the "oven," they are beamed into a vacuum chamber, where they are subjected to excitation by microwave energy. For frequency control, the atoms are made to perform a resonant absorption of energy from the microwave signal; after passing through a second magnetic field, they are deflected toward a hot-wire ionizer. The atoms are then passed through a mass spectrometer that detects and helps remove any contaminants that could otherwise cause random electrical noise bursts. Ion current is converted to electric current by the electron multiplier, and the amplified current is passed through signal processing electronics, which in turn regulates the frequency of a voltage-controlled oscillator. The oscillator's output frequency is multiplied and fed back again to the cesium beam through a waveguide that closes the loop and provides feedback control.

Rubidium vapor standards. Rubidium vapor standards, like cesium beam standards, employ a passive resonator to stabilize a quartz oscillator. Rubidium standards, however, are not self-calibrating and must therefore be checked against a cesium standard during construction. But once built, they are relatively small and are easily transported. Operation (see **fig. 10**,) is based on the principle of the containment of rubidium vapor and an inert buffer gas in a cell illuminated by a beam of precisely filtered light. A photodetector monitors changes, near resonance, in the amount of light absorbed as a function of applied microwave frequencies. The microwave signal is derived by multiplication of the quartz oscillator frequency. Resonance frequency is influenced by the inert buffer gas pressure. This is why rubidium must be calibrated against a cesium reference standard. The Hewlett-Packard model 5065A, shown in **fig. 11**, is one rubidium vapor frequency standard. It exhibits an extremely low drift-rate of less than 10 parts per billion per month and a short-term stability of better than 5 parts per billion averaged over a 1-second period.

time standards

As stated previously, time and frequency are reciprocal quantities; therefore, in order for one to be accurate, its counterpart must likewise be accurate. But how is time precisely defined? This discussion examines several methods and demonstrates the inherent inaccuracies of each.

In 1958, **Ephemeris Time (ET)** based on orbital motion of the earth about the sun at the beginning of the year 1900, was established. But this is a difficult method, and comparisons to ET are impractical. In 1964 the General Conference of Weights and Measures tentatively adopted the "atomic second" based on the number of transitions of the Cesium 133 atom. The standard was fully adopted in 1967.

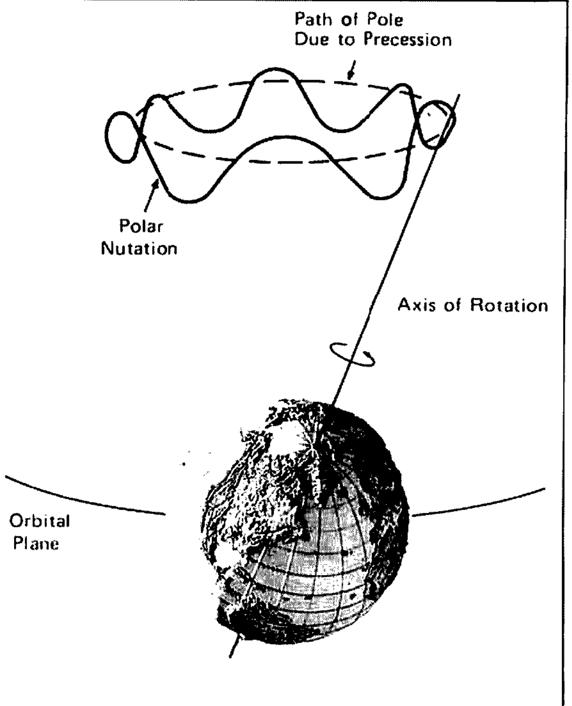

fig. 12. The seasonal wavy motion on the circular path of precession of the Earth's orbit.

table 3. National Laboratories collaborating with BIH.

abbreviation	laboratory	atomic standards qty.	mfr.	type
DHI	Deutsches Hydrographisches Institut, Hamburg, West Germany	1	HP	Cs Std
F	Commission National de l'Heure, Paris, France	11	HP	CS Std
		1	HP	CS Tube with lab electronics
FOA	Research Institute of National Defence, Stockholm, Sweden	2	HP	Cs Std
IEN	Instituto Elettrotecnico Nazionale, Torino, Ittaly	4	HP	Cs Std
IGMA	Instituto Geographico Militar Buenos Aires, Argentina	1	E	Cs Std
ILOM	International Latitude Observatory, Mizusawa, Japan	1	HP	Cs Std
MSO	Mount Stromlo Observatory Canberra, Australia	1	HP	Cs Std
NBS	National Bureau of Standards, Boulder, Colorado	8	HP	Cs Std
		1	LAB	Cs Std
NIS	National Institute for Standards, Cairo, U. A. R.	1	HP	Cs Std
NPL	National Physical Laboratory Teddington, U. K.	4	HP	Cs Std
		1	Lab	Cs Std
		1	Hydrogen Maser	
NPRL	National Physical Research Laboratory, Pretoria, South Africa	1	HP	Cs Std
NRC	National Research Council of Canada, Ottawa, Canada	3	HP	Cs Std
		1	Lab	Cs Std

Apparent solar time is based on the rotation of the earth about its axis with respect to the sun. There are problems with this system because its unit of time derived would be valid only if the sun were to reappear over a fixed point of observation at uniform intervals; of course, it does not. Additionally, there are irregular variations in the rotational speed of the earth and the earth's orbit is elliptical, not circular (see **fig. 12**). The orbital plane, therefore, does not coincide with the plane of the equator. (Then too, the orbital speed of an object whose path describes an ellipse is constantly changing.)

Mean solar time is simply apparent time averaged to eliminate variations due to orbital eccentricity and the tilt of the earth's axis. In a leap year, we should have 365.25 days per year. But actually, we have 365.2444 mean solar days.

Universal time, as with mean solar time, is based on the rotation of the earth about its axis; the units UT were chosen so that on the average, local noon would occur when the sun was on the local meridian. The problem with this system is again that the rotation of the earth is not constant.

Coordinated universal time is corrected universal time which involves the frequency of a precision universal oscillator such as a cesium or rubidium clock being offset from its nominal frequency by an amount which allowed for the clock rate to be nearly coincident with UT_2. (UT_2 is a type of universal time that uses correction factors for seasonal variations in the rotation of the earth.) On January 1, 1972, the UTC system was improved to allow UTC time to accumulate at the same rate as International Atomic Time and therefore eliminate the problem of operating systems adding offsets such as the so-called "leap second" that is added each year.

Laboratories that collaborate with the Bureau International de l'Heure (BIH), the French equivalent of our NBS (see **fig. 13**), are listed in **table 3**. In addition to the three hydrogen masers listed, it is likely that the U.S.S.R. has at least one maser, although this information is hard to obtain.

There are other systems, including apparent sidereal time, and mean sidereal time, but it is sufficient to say that these herculean attempts to establish accurate time intervals are made so that somebody can have an oscillator that is *truly* fine-tuned. Seriously though, if this area of time-keeping seems fascinating to you, as it does to me, then you may want to obtain map No. 76, depicting worldwide time zones, from the U. S. Navy Defense Mapping Agency.[1] You may also wish to be placed on the mailing list of the *NBS Time and Frequency Services Bulletin* issued monthly by the Time and Frequency Division of the National Bureau of Standards.[2]

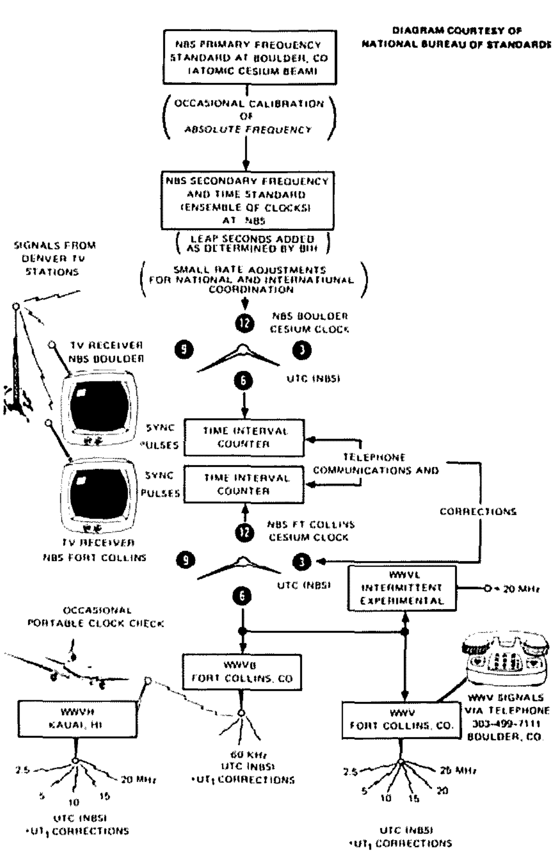

fig. 13. National Bureau of Standards frequency and time facilities.

acknowledgments

I would like to thank P. K. Weir of Hewlett-Packard and Lupe Lopez and Dick Rogers of John Fluke Mfg. and their respective companies as well as Tracor, Inc., for providing technical assistance and artwork for this article.

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1. Address requests to MC2 Sales, U. S. Navy Defense Mapping Agency, Hydrographic Center, Washington, D. C. 20390. Enclose check or money order for $5.00.
2. Frequency-Time Broadcast Services Center, Time and Frequency Division, National Bureau of Standards, Boulder, Colorado 80302.

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Sprang, Steve, *Calibration-Philosophy in Practice*, Application Note, John Fluke Manufacturing Co., Inc.

Fundamentals of Time and Frequency Standards, Hewlett-Packard Application Note 52-1, October, 1974.

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ham radio

Combine a scope and an NE555 multivibrator to build an important diagnostic tool

a time domain reflectometer

Before I describe how to build a TDR and discuss how it works, consider this simple example of the theory behind time domain reflectometry. Assume you're standing some distance away from the front of a large building. Yell, and you'll hear your echo. If you know the speed of sound and measure the time it takes between your initial call and the return of your echo, you can determine the distance between yourself and the building. Just multiply the speed of sound by the time it took for you to first hear your echo and divide by two. (You divide by two because the sound wave goes to *and* from the building; all you need to know is the distance in one direction.)

Suppose, now, that you get a call from the local repeater group saying the repeater cannot be heard. They believe the malfunction is in either the transmission line or the antenna, but they're not certain which it might be. With a time domain reflectometer, you can tell whether the problem is in the antenna or the line — and if the trouble is in the line, approximately where it is.

By sending a pulse or transition of levels down a transmission line, and then observing the reflected signal, you can determine whether the transmission line is open, shorted, or terminated by some value of resistance. If the termination resistance is different from the characteristic impedance of the line, its value can also be approximately determined.

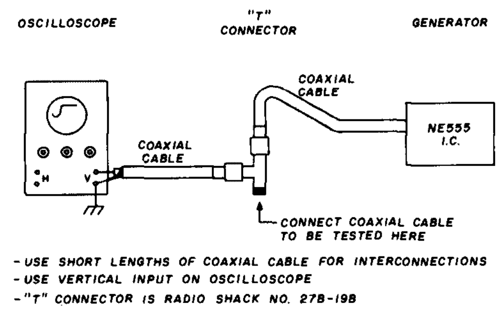


fig. 1. Equipment needed to develop a time domain reflectometer.

setting up

The test setup is illustrated in **fig. 1**.

The more accurately you measure the time, the more accurately you'll be able to locate the fault in the transmission line or antenna. Consequently, your oscilloscope should be capable of measuring time periods down to 0.1 μS and have a reasonably accurate time base. The bandwidth of the scope is not critical, but should be at least 5 MHz. Vertical sensitivity is relatively unimportant; any scope capable of measuring video signals should be adequate.

By Bill Unger, VE3EFC, 431 North Syndicate, Thunder Bay, Ontario, Canada P7C 3W9

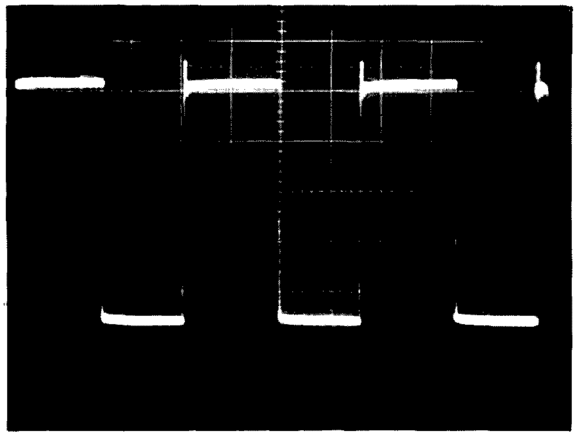


fig. 2. Output of an NE555 multivibrator. Horizontal deflection is 5 microseconds per centimeter.

Some scopes have a terminal marked "Gate Output," a pulse that coincides with the sweep time; this can also be used to provide the signal. It is important, however, that the scope output have the same impedance as the line you want to measure.

A 555 IC wired as an astable multivibrator provides the pulse train as it oscillates at a frequency of 60 kHz (see **fig. 2**). Frequency here is not critical; if yours is slightly different, that's fine. The value of R1 in the circuit should match the impedance of the line being tested — generally 50 or 75 ohms (**fig. 3**).

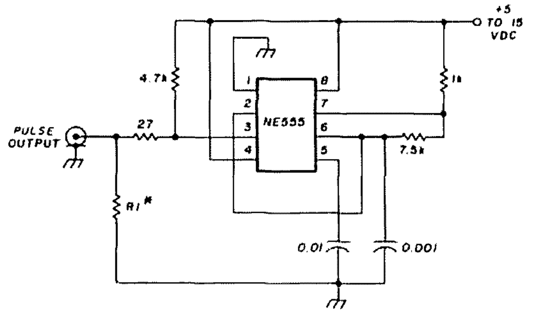
This entire assembly can be built on a Radio Shack project board (No. 276-024), with a switch added for selection of either 50 or 75 ohms output.

interpreting results

If the line is properly terminated in its characteristic impedance, the trace will appear as a straight line, indicating that all of the forward signal has been dissipated in the resistance and no reflection exists (see **fig. 4**). If, however, there is a reflection and the trace goes up, this would indicate that the end of the line is either open or of higher impedance than the cable (see **fig. 5**). If the trace goes down, the line is either shorted or of lower impedance than the cable (see **fig. 6**). As an aid to remembering which is which, consider this: when measuring voltage, a short (or low impedance) lowers the voltage and an open allows the voltage to rise.

To determine the location of the fault, you must measure the time between the initial pulse and the pulse caused by the problem. The speed of radio waves in free space is 983.5 feet per micro-second, so if you multiply the time it takes by the speed of light, the result will be the distance to the fault in feet. However, because the radio wave is slowed by the velocity factor of the coax, you must therefore multiply the previous distance by the velocity factor and then divide by two to determine location of the malfunction.

Perhaps the method will be clearer if we look at an example. Suppose we have a piece of RG-213, with a velocity factor of 0.66. A problem exists at a point that measures 0.3 μS down the line. The distance to the malfunction is the speed of radio waves multiplied by the time (983.5 $\times$ 0.3) = 295 feet. Correc-



* R1 = IMPEDANCE OF LINE UNDER TEST - EITHER 50 OR 75 OHMS.

Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

fig. 3. Schematic diagram of an NE555 astable multivibrator used in the time domain reflectometer.

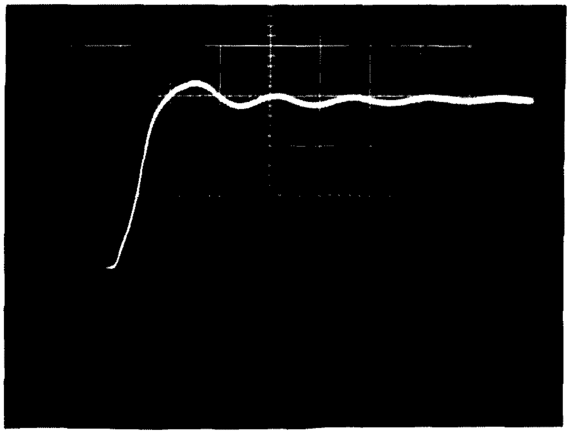


fig. 4. One hundred feet of RG-213 transmission line is terminated in 50 ohms. Note the basically flat response after the pulse arrives.

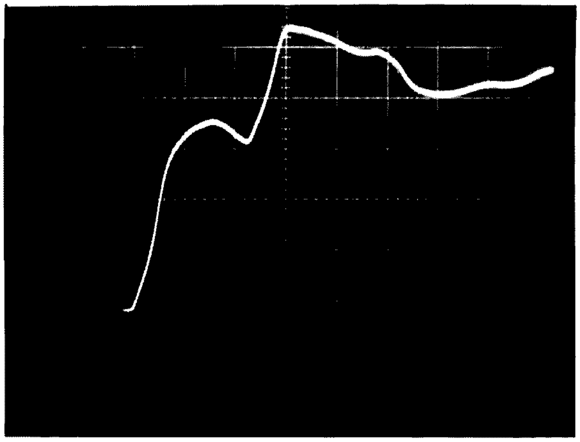


fig. 5. One hundred feet of RG-213 is now pulsed while an open exists at the far end (load end). The rapid rise in pulse height marks the location of the "fault" (open circuit). Horizontal deflection is 0.1 microseconds per centimeter.

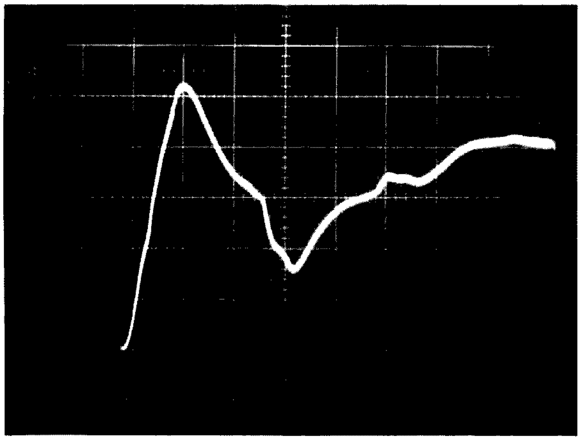


fig. 6. The same length of RG-213 is pulsed while a short exists at the free end. The rapid drop in pulse height marks the location of the "fault" (short circuit). Horizontal deflection is 0.1 microseconds per centimeter.

ting for the velocity factor, we multiply 295 feet by 0.66 to obtain 194 feet. Since this is the total distance the signal traveled, dividing 194 by two determines the distance from the measuring point to the fault. In this case the distance is 97 feet.

I have prepared a graph (fig. 7) showing the distance in feet versus the time to the problem, on which the result can be read directly in feet. There are two lines, each representing a different velocity factor; the top line represents transmission lines with a velocity factor of 0.80, which is typical of foam-filled lines, and the bottom one, a velocity factor of 0.66, which would include coax such as RG-8, RG-58, and RG-213.

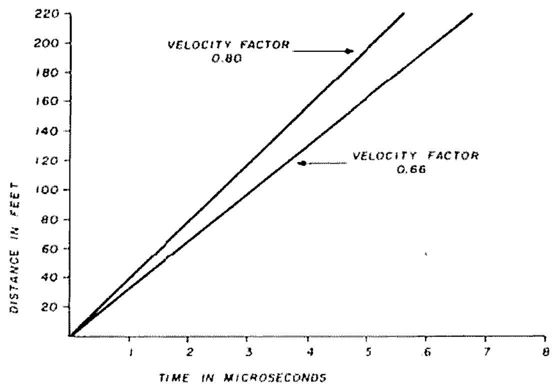


fig. 7. Chart enables rapid determination of the distance to a "fault" if elapsed time is known. Two cables with different velocity factors are considered.

conclusion

If your oscilloscope doesn't have a calibrated time base you may still be able to employ this method by using a section of cable of a known length and then expressing the unknown cable in lengths of the known one. If a 100-foot cable takes 0.3 μS to indicate a problem, then a measurement of 0.6 μS would indicate a cable 200-feet long. (Be sure the cables you use have the same velocity factor; if they don't, the comparison will not be valid.)*

If you want to measure cables that are less than 40 feet in length, the task becomes rather difficult because of the extremely short times involved. In this case it would be desirable to add a 100-foot section of coax and then subtract 100 feet when you are calculating the distance.

Now back to that phone call from the repeater group. By taking the TDR you've built from a scope and an NE555 multivibrator to the base of the tower — and keeping the principles of time domain reflectometry in mind — you'll be able to tell the climbers roughly where and what the problem is.

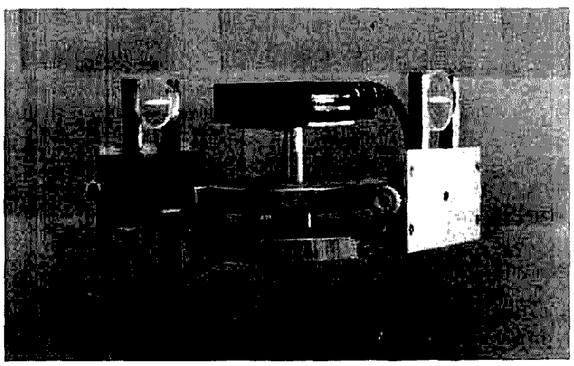
ham radio

*Even if the velocity factors differ, the fault can still be located as long as the ratio of velocity factors is known. While this method cannot be used to measure complex impedances, it is useful in determining problems along the line. Keep in mind that the antenna at the end of the line may be either open or shorted; this wouldn't be a fault on the line itself. **Editor**

construct an optical FM receiver

"See" the entire FM band and detect each station independently and simultaneously

The ability to see — instantaneously — and detect a wide band of frequencies is possible using an electro-optical technique known as Bragg cell operation.[1,2] A receiver providing this performance consists of an acoustic medium such as a slab of glass, a transducer bonded to the glass, a light source, optical elements and associated RF drive circuits.[3] What is achieved is a series of position-modulated light spots in the output (plane), each representing a separate station with its own information content. The practical result of this device is an FM receiver achieving 200 kHz resolution, sufficient to separate and detect all FM band stations.



Low-light photograph shows lateral width of light beam traversing Bragg cell.

Bragg cell operation can be understood in terms of a few physical concepts. An RF signal when applied across the terminals is transformed into a traveling acoustic (sound) wave in the cell. The pressure in the sound wave creates a traveling wave of rarefaction and compression in the glass medium which in turn causes analogous changes in the index of refraction. The Bragg cell, as shown in **fig. 1**, acts as a phase diffraction grating with an effective grating line separation equal to the wavelength of sound Λ (which is related to the wavelength of the RF-injected signal). The deflection angle ϕ_d (measured outside the cell) between the incident and the first order light term equals

$\frac{\lambda_o}{\lambda} = (\lambda_o \bullet f/V_s)$ where f is the frequency and V_s the sound velocity.

Incident light is most efficiently diffracted when the incident angle equals $1/2 \bullet \frac{\lambda_c}{\lambda}$ where λ_c refers to the center of the band. This angle, ϕ_B, known as the Bragg angle, may also be written as:

$$\phi_B = \lambda_o f_c / 2V_s \qquad (1)$$

where f_c is the center frequency, *and*
V_s is the sound velocity

This design uses a Bragg cell made of glass (V_s = *4000* m/second and the light source is a He-Ne laser [λ_o = *6328Å* (Angstrom unit)]. Substituting these values in eq. **1** shows that the Bragg angle is equal to:

$$\phi_B = 7.91 \times 10^{-5} f_c\,(MHz)\ radians,\ or$$

$$\phi_B = 4.53 \times 10^{-3} f_c\,(MHz)\ degrees \qquad (2)$$

Denoting the incident angle by ϕ_B, the Bragg cell operation can be represented as shown in **fig. 1**. All

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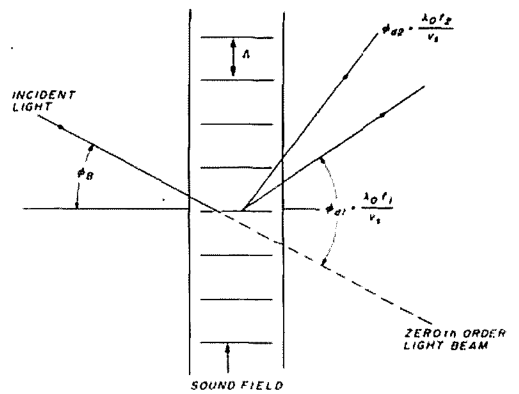


fig. 1. Bragg cell shows relationship between incident and diffracted light rays.

angles have been exaggerated for clarity. The Bragg cell diffracts light rays into angles controlled by spectrum of acoustic frequencies f_i, where $i = 1, 2$, etc. This acoustic spectrum is identical to the frequency spectrum of the electrical signal. Though Bragg cells exhibit limited bandwidth, this design is sufficiently wide to accommodate the entire FM band.

system configuration

By definition, an electro-optical receiver contains both electronic circuits and optics (see **fig. 2**). The electronics are used as the input and output stages of the system with an optical medium in between. The input stages "condition" the received signal, providing compatibility with and driving the optics interface. The input stage electronics consists of an FM low-level amplifier, balanced mixer, local oscillator and RF power amplifier. Specifically an FM signal is amplified in the low-level stage (using a Radio Shack FM amplifier), mixed with a local oscillator in the HP 10514A mixer, and further amplified by the RF power amplifier (Intra-Action EE-40), achieving a power level of 1-2 watts necessary to drive the Bragg cell. Since the Bragg cell (Intra-Action ADM-40) is tuned for operation at 40 MHz, with an effective bandwidth of 40 MHz, it is necessary to translate the FM band (center frequency 97 MHz) down to this center frequency. This is accomplished by mixing with a 57 MHz local oscillator. In addition, the associated Bragg angle for 40 MHz, according to **eq. 2**, is 0.181 degrees.

The optics, shown in **fig. 2**, include four lenses, two of which are identical spherical converging lenses (L2 and L3, having a focal length of F2 ≅ 20 cm) and two are identical cylindrical lenses (L1 and L4, having a focal length of F1 ≅ 3 cm). The optical processing which precedes the Bragg cell is intended to spread the beam laterally (i.e., in the plane of the paper) which, for reasons to be explained later, en-

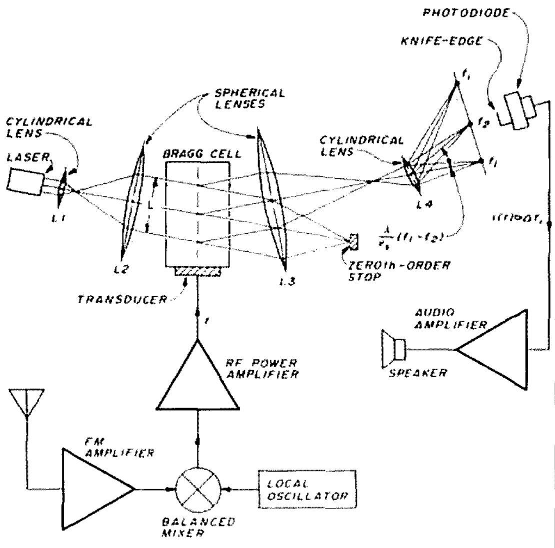
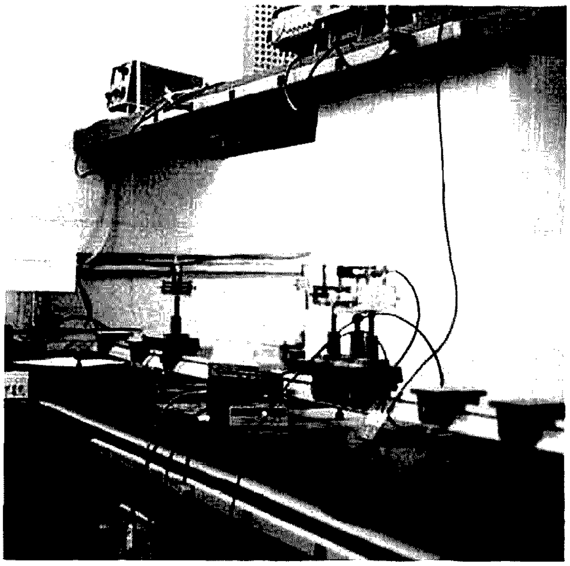
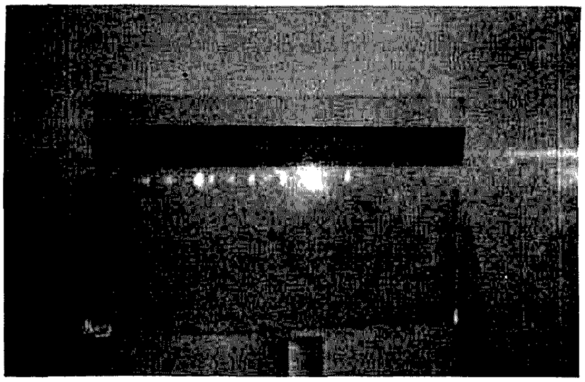


fig. 2. FM optical receiver block diagram.



Completed optical FM receiver.



Separated laser spots, each representing a different FM station. Large spot at right is zeroth order beam, schematically represented in *fig. 1*.

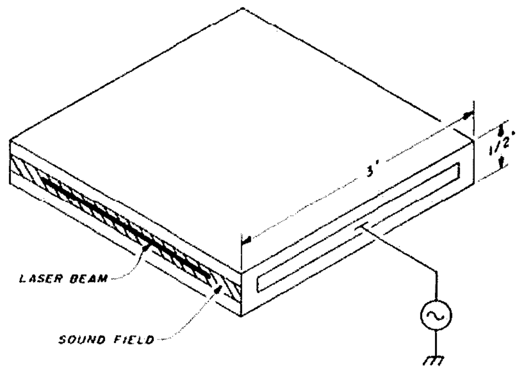


fig. 3. Acousto-optic Bragg cell with laser beam and sound field illustrated for clarity.

hances the frequency resolution. This situation is shown in fig. 3, and in the photograph above. The sound power is concentated in the cell along the center with a height of approximately 2 mm. Consequently, in order not to throw power away, the beam has been expanded in one direction only; this is accomplished by using a cylindrical lens for L1 instead of a spherical lens. The laterally diverging beam which follows L1 is collimated before entering the Bragg cell with a spherical lens L2. L2 is placed a distance equal to the sum of the focal lengths, F1 + F2, away from the cylindrical lens L1 so as to form a telescope configuration. A beam of fixed lateral width L, which is controlled by the ratio of F2/F1, is incident on the cell at the Bragg angle as shown in fig. 2.

Each FM carrier has an independent light beam diffracted in a direction determined by the carrier frequency. For clarity, only a few of the diffracted light beams are shown in **fig. 2**. The second spherical lens, L3, is intended to focus the emerging beams in its back focal (output) plane. The second converging cylindrical lens, L4, is placed in such a way that the back focal plane of L3 is imaged in the plane of the knife edge in front of the photodiode (see **figs. 2** and **4**).

theory of operation

For the i^th^ FM station the signal's instantaneous frequency is represented by

$$f_{FM_i} = f^{o}_{FM_i} + \Delta f_i(t) \tag{3}$$

which is the sum of a fixed carrier frequency $f^{o}_{FM_i}$ and a time varying frequency difference $\Delta f_i(t)$, the latter being proportional to the audio signal. The FM variation Δf_i is small compared to the carrier $f^{o}_{FM_i}$. Using **eq. 3** the i^th^ FM station is beamed, on the average, in a direction given by

$$\phi_{di} = (\lambda f_i / V_s)$$
$$\cong 1.58 \times 10^{-4} [f^{o}_{FM_i}(MHz) - f_m (MHz)]\ radians,$$
$$or \tag{4}$$
$$\cong 9.06 \times 10^{-3} [f^{o}_{FM_i}(MHz) - f_m (MHz)]\ degrees,$$

where f_m is the mixing frequency (57 MHz). This is illustrated in fig. **5**. The actual instantaneous angle of deflection deviates slightly from the above angle due to the inclusion of $\Delta f_i(t)$ which causes a 'wobble,' $\Delta\phi_{di}$, in the deflected beam:

$$\Delta\phi_{di} = 1.58 \times 10^{-4} \Delta f_i (MHz)\ radians\ or$$
$$= 9.06 \times 10^{-3} \Delta f_i (MHz)\ degrees. \tag{5}$$

This relationship between the audio signal (as encoded in Δf_i) and the variation in the deflected angle is used to generate an electrical signal with an *amplitude* proportional to the strength of the signal. By placing a knife edge screen in front of a photodiode (see fig. **6**) in the detection plane, the integrated light

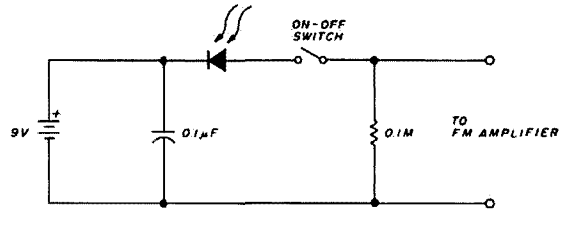


fig. 4. Photodiode circuit used in FM receiver.

intensity varies with the *wobble*, $\Delta\phi_{di}$, and hence provides a current proportional to Δf_i, i.e., proportional to the audio signal.

frequency resolution

Resolution is proportional to the lateral width of the light beam transversing the Bragg cell. The number of resolvable angles[2] for a total frequency change, Δf, is given by:

$$N = (d/V_s)\Delta f \tag{6}$$

where d is the lateral width of the light beam. You may notice that (d/V_s) is the transit time of the sound as it transverses the light beam. In our case, the laser beam width is increased from 1 mm to 20 mm (2 cm), which is the lateral width L as shown in **fig. 2**. Since the full FM band (88-108 MHz) is used, Δf is then 20 MHz. Substituting the velocity of sound for glass, 4×10^3 m/sec in the equation, the calculated transit time, (0.02 m/4×10^3 m/sec), is 5×10^{-6} second. Direct substitution into **eq.** 6 results in 100 resolvable points over the FM band or a frequency resolution of 200 kHz. Therefore all FM stations are resolvable. **Eq.** 6 can be used to predict the resolution for any Bragg spectrum analyzer application.

In summary, the audio signal of the ith FM station, Δf_i, results in a wobbling diffracted beam, $\Delta\phi_{di}$, which is then transferred to the electrical domain by using a photodiode positioned in the shadow of a knife edge.

where to get parts

Electronics obtainable from Radio Shack are:

Archer FM amplifier, Catalog No. 15-1122

Realistic stereo integrated-amplifier SA-102, Catalog No. 31-1963

FM antenna, Catalog No. 15-1639

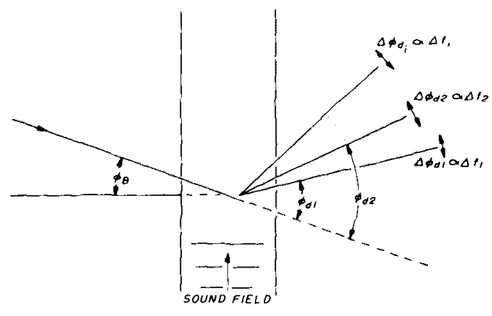


fig. 5. Diffracted beam paths that include audio signal modulation or "wobble."

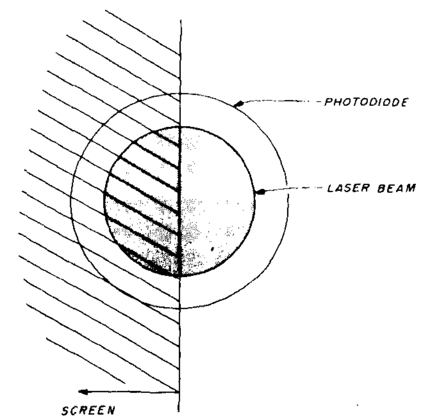


fig. 6. Knife-edge technique for the detection of wobble in diffracted light beam.

Archer color matching transformer, Catalog No. 15-1140

Realistic Minimus speaker, Catalog No. 40-1996

The following optics are available from Edmund Scientific Co., 101 E. Gloucester Pike, Barrington, New Jersey 08007:

He-Ne laser

Spherical converging lens (F ≅ 20)

Cylindrical converging lens (F ≅ 3)

The following are available from Intra-Action Co., 3719 Warren Avenue, Bellwood, Illinois 60104, telephone 312-595-3770:

RF amplifier — E-40

Bragg cell — ADM-40 or AOM-40

acknowledgments

We would like to thank Professor A. Korpel for his suggestions and critical review. The device discussed here represents an educational aspect of more fundamental research into acousto-optics supported by the National Science Foundation under grant #ECS-8121781.

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ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

I met Pat Hawker, G3VA, some years ago at an electronics trade show in New York, and since then, we've QSO'd from time to time. One of Pat's attributes is his ability to ferret out new ideas that otherwise could be lost in the noise of the day-to-day progress of the electronics world. I always enjoy reading Pat's column, "Technical Topics," in the magazine Radio Communication, *the flagship publication of the Radio Society of Great Britain.*

"absorbing yagi"

In his November, 1982, column, Pat described the "absorbing Yagi" antenna scheme of John Beech, G8SEQ. It appears that John has developed a new technique that improves the pattern of the conventional Yagi antenna, particularly in regard to the front-to-back ratio of this popular antenna (**fig. 1**). It's not difficult to achieve good gain with a Yagi; the antenna is most forgiving when it comes to adjustment and layout. The adjustment of front-to-back ratio, on the other hand, is both sensitive and crucial, and will vary greatly from one installation to the next. G8SEQ has achieved very high front-to-back ratios by the addition of a new element to the Yagi, which he calls an *absorber element*. He reports ratios as high as *75* dB for a 13-element VHF array!

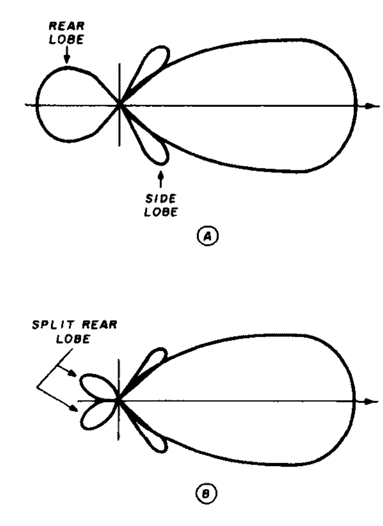

fig. 1. (*A*) Representation of field plot of a conventional Yagi showing side and rear lobes, and (*B*) Plot of "absorbing Yagi" showing reduction of unwanted lobes. The deep null to the rear of the array can be maximized by "fine tuning" the absorber element.

As the name suggests, the absorber element absorbs energy that would otherwise be radiated to the rear of the array. In its simplest form it is an extra dipole element, resonant at the operating frequency of the Yagi, with a resistor placed at its center. The resistance is approximately equal to the center impedance of the driven element (**fig. 2**).

The absorber adjustments consist of varying the spacing to the reflector and adjusting the value of the center resistor until optimum front-to-back ratio is observed. If the Yagi has a reasonable front-to-back ratio to begin with, an absorber resistor power rating of 25 watts will suffice even at full legal power.

According to Hawker, "G8SEQ suggests that one can regard the Yagi array as a directional bandpass filter, and that the absorber is the element that has been missing for years. With the same thinking now sometimes being applied to absorptive lowpass TVI filters, the unwanted RF is safely dissipated in a dummy load rather than attempting merely to 'short circuit' it with a reflector."[1]

While no specific dimensions are given, G8SEQ suggests that the absorber element be self-resonant at the operating frequency and placed about 0.23 wavelength behind the re-

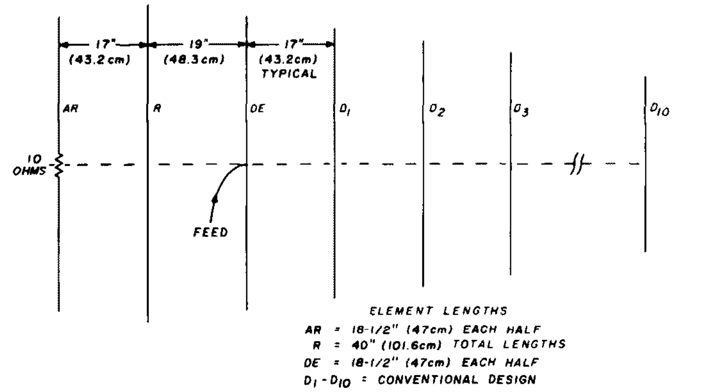

fig. 2. The G8SEQ absorbing Yagi for 144 MHz. The 12-element beam has conventional dimensions plus the addition of an absorber element 17 inches behind the normal reflector. Absorber has same dimensions as driven element. Elements are made of 3/8-inch diameter tubing.

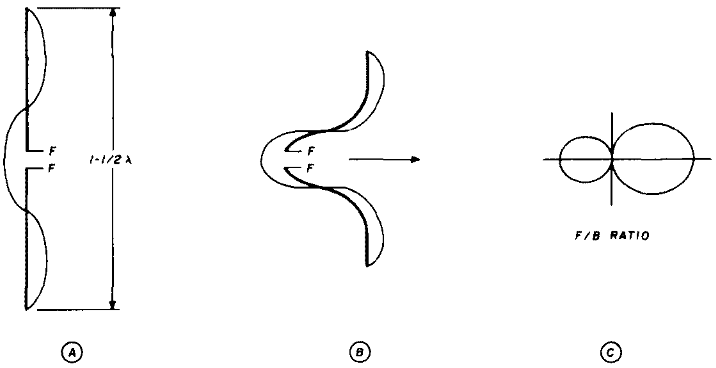

fig. 3. (*A*) Representation of current distribution on 3/2-wavelength dipole, and (*B*) optimum-shaped dipole. (*C*) Current distribution in modified dipole causes radiation to increase in forward direction.

flector and that the center resistor be about 10 ohms, noninductive. I would guess that less spacing could be used, provided absorber length and resistance were varied.

I think this is a good idea and I would be pleased to hear from any experimenters who try this novel technique. In today's world of heavy interference, antenna front-to-back ratio may be of more importance than antenna gain.

optimum-shaped antenna element

For some time I have heard about a new antenna element design that provided increased gain and improved operating characteristics. Again, Pat Hawker tracked it down and described it in his November column. The antenna, discussed at an International Conference on Antennas and Propagation in London in 1979, was developed by F. M. Landstorfer at the Technical University in Munich. Further work was done on the antenna by Cheng and Liang of Syracuse University in 1982.

For many years it was assumed that the "building block" of a beam antenna was a straight dipole element about a half-wavelength long. Landstorfer investigated the use of 3/2-wavelength element, curved in a specific manner, to provide forward gain and directivity (**fig. 3**). Landstorfer's element has a forward gain of about 5 dBi.

The main disadvantage of this idea is simply the larger size of the basic element. Even so, this is compensated for by the fact that far fewer elements are needed to obtain equivalent gain.

A three-element Yagi using this technique (**fig. 4**) was tested. The gain was measured at 11.5 dBi (about 9.4 dB over a dipole), with a front-to-back ratio of 26 dB. Sidelobe attenuation was better than 20 dB.

G3VA points out that the optimized shape of the elements is related to element diameter, and until more specific information is available, the cut-and-try technique is recommended for those wishing to experiment with this novel antenna.

the terminated, traveling-wave antenna

A final note before we leave Pat Hawker, G3VA. Pat wrote about an interesting antenna development originally described in the *IEEE Transactions on Antenna and Propagation*,[2] by Matsuzuka and Nagasawa of Nihon University (home of the famous Dr. Yagi of Yagi *antenna* fame), Tokusuda, Japan. Matsuzuka

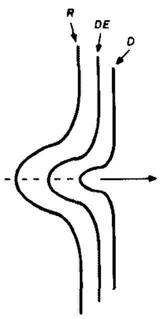

fig. 4. Top view of experimental Yagi using gain-optimized elements, each approximately 3/2-wavelengths long. VHF array of this design provided 11.5 dBi gain and a front-to-back ratio of about 26 dB.

and Nagasawa described a rectangular loop antenna, fed at the center of one side and terminated with a resistor placed in the opposite side (**fig. 5**). The loop provides a unidirectional pattern with the approximate dimensions shown. A very high (unspecified) front-to-back ratio is achieved over a significant bandwidth. Dimensions and resistor value are relatively non-critical.

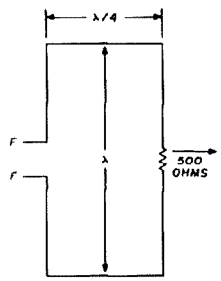

fig. 5. The terminated, traveling wave loop antenna provides good directivity, front-to-back ratio and bandwidth. Value of terminating resistor is not critical. RF energy reaching far end of antenna is dissipated in the resistor rather than being reflected back to the feedpoint. Wattage rating of resistor is equal to about one-half the power output of the transmitter.

VCR RFI: more problems for hams

I've heard that some Amateurs are experiencing RFI problems with their transmissions interfering with video cassette recorders (VCRs). I have a VCR myself, and have had no problems with it, perhaps because I'm not on the air when I'm watching it. However, I would appreciate hearing from any readers who have had VCR RFI and solved the problem (if they did). I have also heard that video disc recorders are RFI-prone, especially to signal in the region of the forthcoming 800 MHz Amateur band and the forthcoming mobile communication band.

As solid-state electronics invades our lives more and more, the RFI problem will become more severe. Unfortunately, even though Public Law 97-256 requires manufacturers to produce RFI free equipment, the FCC favors voluntary standards, rather than imposed standards. This leaves the door wide open to abuse and circumvention of the law, leaving the Radio Amateur to take the blame for RFI caused by poorly designed products.

the 4-1000A linear amplifier

There's a considerable amount of interest in using the 4-1000A tetrode as a cathode-driven, linear amplifier, and it is a popular tube for "home brew" equipment. In grounded grid service, the tube will operate well at plate potentials between 3 and 5 kV and can easily provide up to the legal 1.5 kW PEP output limit. Typical operation of this tube at 3 kV is as follows:

DC plate voltage	3 kV
zero-signal plate current	55 mA
single-tone plate current	700 mA
single-tone grid current	275 mA
single-tone driving power	120 watts
load impedance	2350 ohms
driving impedance	104 ohms
plate input power	2100 watts (PEP)
useful output power	1320 watts (PEP)
3rd order distortion	−34 dB
5th order distortion	−36 dB

Note: Single-tone grid current is sum of grid one and grid two currents. Useful power output is power delivered to the load. Plate power output is 1500 watts PEP.

For full information on the 4-1000A and other glass tubes suited for grounded grid service, write to me at EIMAC, 301 Industrial Way, San Carlos, California 94070. (Enclose two first-class stamps or two IRCs for postage.)

Circuit design and additional information on linear amplifiers may be found in the 22nd edition of *The Radio Handbook*, published by Howard W. Sams and available from Ham Radio's Bookstore, Greenville, New Hampshire 03048.

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ham radio

EMI/RFI test receivers

Specialized receivers demand tighter design requirements

> This article was prepared from a paper originally delivered at Electro 82, an electronic show and convention held in Boston on May 25-27, 1982. **Editor**

Many electrical devices — from computers to hair dryers — generate electrical noise. This noise is either conducted along power cables to other electrical devices, or radiated through the unit's enclosure, keyboard, or screen to the world outside, thereby producing interference that affects still other electrical instruments.

A receiver able to detect and measure this interference and relate it to precise (accepted) international standards must necessarily be based upon a different design than that of a communications receiver. The latter type normally covers the 10 kHz to 1000 MHz frequency range and is primarily designed to receive and decode selected transmissions. Conversely, an EMI receiver must include the following design features:

A greater instantaneous dynamic range than that of a communications receiver, since the energy of incoming pulses can be higher than several intelligence-bearing signals or constant carriers

A circuit that monitors the maximum allowable voltage at different stages, to prevent short-term overload

Detector time constants that conform to internationally agreed-upon standards

Appropriate IF bandwidths

Precise amplitude calibration over the operating frequency range (Attainment of this normally requires the use of either a spectrum generator or tracking generator.)

If active antennas are used to measure the field in the low frequency range, they must have the *necessary dynamic range and proper antenna correction factor.*

The different requirements of test receivers and normal communications receivers will be discussed in this article, with special attention paid to the relative advantages or disadvantages of manual and automatic measuring capability.

dynamic requirements

Before specific EMI/RFI receivers — such as the Rohde & Schwartz ESH2 manually-operated

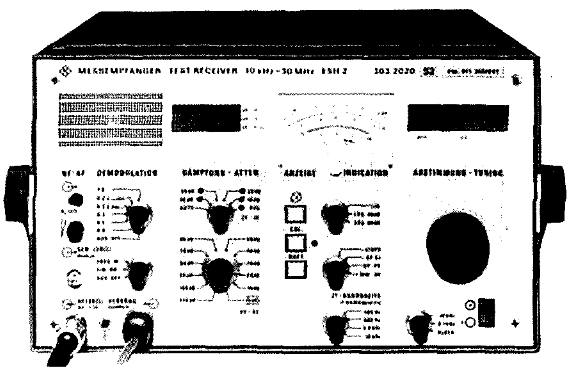


fig. 1. Manually operated EMI/RFI test receiver, type ESH. Frequency range, 10 kHz to 30 MHz.

By Dr. Ulrich L. Rohde, DJ2LR, President, Communications Consulting Corp., 52 Hillcrest Drive, Upper Saddle River, New Jersey 07458

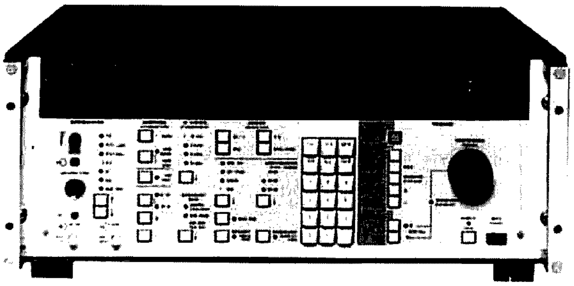


fig. 2. Microprocessor-controlled EMI/RFI test receiver, type ESH3, features built-in intelligence. Frequency range, 10 kHz to 30 MHz.

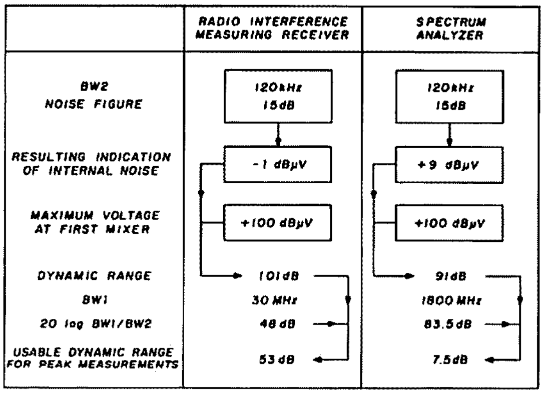


fig. 3. Comparison of EMI/RFI receiver and spectrum analyzer to evaluate usable dynamic range.

EMI/RFI 10 kHz to 30 MHz receiver (**fig. 1**) or the ESH3 computer-controllable EMI/RFI test receiver with built-in intelligence (**fig. 2**) — were introduced, spectrum analyzers were generally used to detect and characterize emitted noise spectrums. There has been some controversy as to whether spectrum analyzers that employ special "quasi-peak" detectors (CISPR/ANSI) can provide the necessary information. This is an important issue and should be clarified.[1]

The spectrum analyzer, while quite capable of rapidly providing data on CW and various sinusoidal signals, is not as suited to measure pulse spectrum parameters with the same facility. To understand why, a discussion on spectrum and bandwidth requirements is called for.

Electrical pulses of short duration possess considerable energy over a wide frequency range. When this signal is introduced into a bandpass filter, the output peak voltage (of the pulse) is proportional to the pulse bandwidth (which is approximately the 6 dB bandwidth of the filter).

$$E_{peak} \propto BW\ (6\ dB) \qquad (1)$$

If a signal having a pulse spectrum is introduced into two cascaded bandpass filters with different bandwidths, $BW1$ and $BW2$ (with $BW1 > BW2$), the ratio of the output peak voltage is equal to the ratio of the filter bandwidths or:

$$\frac{E_{1peak}}{E_{2peak}} = \frac{BW1}{BW2} \text{ or,} \qquad (2a)$$

$$\Delta E\ (dB) = 20 \log \frac{BW1}{BW2}\ dB \qquad (2b)$$

significance of different bandwidths

Here, the question of RF preselection (input RF bandwidth) comes into play. If no preselection exists (as in the case with a spectrum analyzer), the measured output levels (analyzer and receiver) are different. Assume you are testing in the 30-1000 MHz range. Typical input filter bandwidths are

measuring receiver $BW1 = 30\ MHz$
spectrum analyzer $BW1 = 1800\ MHz$

Consequently, the voltage E1 presented to the first mixer of the device is 48 dB and 83.5 dB higher, respectively, than the output (indicated) voltage E2. Therefore, the narrower RF filter of the measuring receiver lowers the required mixer dynamic range by 35.5 dB (83.5 − 48 = 35.5).

Let us apply these facts to the measuring receiver and spectrum analyzer. We have assumed that each device uses the same mixer (with equal maximum input voltages), and the receiver has an approximately 10 dB lower noise figure. (This is typical, though the difference may even be greater, as in **fig. 3**.)

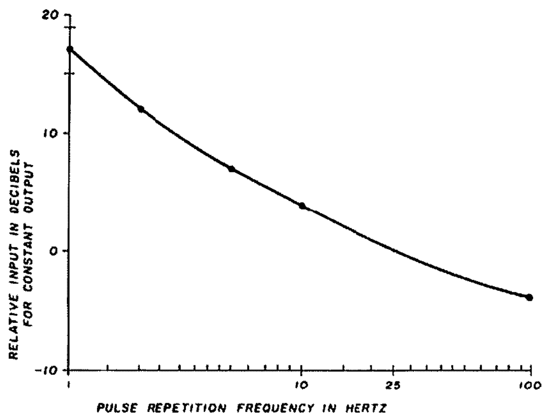


fig. 4. Pulse response curve for the frequency range 10 kHz to 150 kHz (per CISPR Publication No. 16, 1977).

Therefore, one can see that when making "peak" measurements with a spectrum analyzer the usable dynamic range is so limited that the measurement must be monitored carefully to assure the linear operation of the mixer. This can be accomplished by switching in a small amount of attenuation and comparing this value to the drop in measured output. However, this is time-consuming, and definitely not in line with the requirements for rapid automated testing.

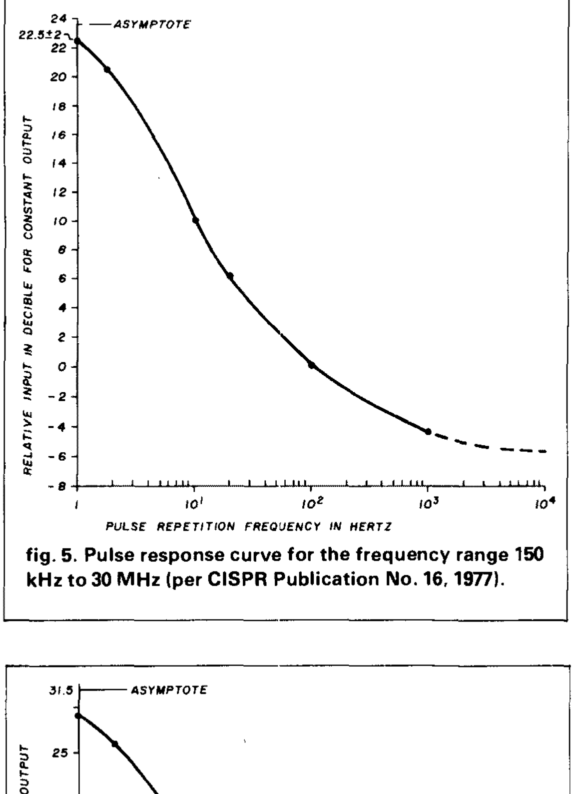


fig. 5. Pulse response curve for the frequency range 150 kHz to 30 MHz (per CISPR Publication No. 16, 1977).

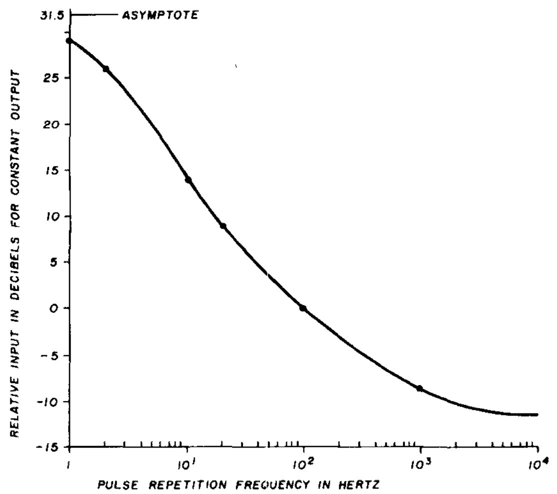


fig. 6. Pulse response curve for the frequency range 30 MHz to 1000 MHz (per CISPR Publication No. 16, 1977).

problem with quasi-peak detectors

The most serious flaw in the application of spectrum analyzers is in the use of a "quasi-peak" detector. A "quasi-peak" detector is simply a weighting network that gives a weighted indication based on the PRF (pulse repetition frequency) of the incoming pulse spectrum. (The curves for this weighting are shown in **figs. 4**, **5**, and **6**). The variation in weighting in the VHF/UHF range is 39.5 dB. It is impossible for a measuring device with a usable dynamic range of only 7.5 dB to give a correctly weighted output over a 39.5 dB range. (Remember, the weighting circuitry is at the IF, **after** the "damage" is done.)

The final conclusion is that, based on the simple physics of the measurement, it is difficult to measure pulse spectra peaks with a spectrum analyzer, and the use of "quasi-peak" circuitry at the IF of an analyzer is impossible without appropriate RF preselection. (**Fig. 7** shows overall selectivity as a function of frequency range required for the EMI/RFI test receiver to meet specifications.)

high-dynamic range required

Fig. 8 shows the block diagram of a modern RFI test receiver. It consists of an RF attenuator, a built-in calibrator, a tracking input filter, a mixer, IF stages, and the detector for demodulation, as well as the required weighting filter and rectifiers.

It becomes immediately apparent that the major difference between this block diagram and the block diagram of a typical communication receiver is the RF attenuator, the calibrator, and the lack of preamplification ahead of the mixer.

Assume that a high level double-balanced mixer is used, and that both the RF attenuator and the bandpass filter do not introduce any intermodulation distortion. In this case, the large signal performance of the receiver is determined by the mixer and the stage immediately following the mixer, most likely a termination amplifier with a crystal filter immediately following it.

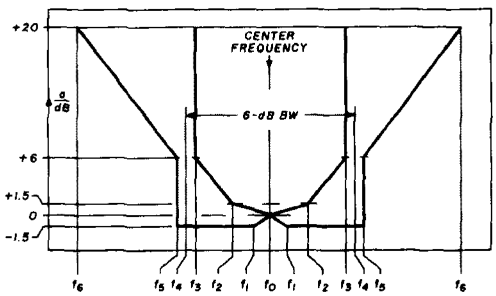
The mixer, typically a passive device, introduces 5.5 to 6 dB of loss to the next stage (an amplifier). Most likely, these two stages determine the overall intermodulation distortion performance of the receiver. The high level double balanced mixer and the post-amplifier probably have a +30 dBm intercept point.

The presence of the input filter not only reduces the number of signals but also improves the second order intermodulation distortion substantially, relative to a wide-open front end.

prevention of overload

The receiver can saturate if the combined signal level present at the output of the input tracking and IF crystal filters is excessive. While it may not be possible to prevent such an overload condition initially, it is important to detect the condition. The input RF attenuator can then be used to reduce the overload.

The automatic and computer controllable EMI/RFI receiver ESH3 *automatically switches in the required* attenuation to make sure that this overload condition does not occur while providing a 60 dB dynamic range at the IF.



RF RANGE MHz					
	MIN	0.010	0.15	30	MHz
	MAX	0.150	30	1000	MHz
$f_1 = f_0 \pm$		0.05	1	10	kHz
$f_2 = f_0 \pm$		0.05	2	20	kHz
$f_3 = f_0 \pm$		0.085	4	50	kHz
$f_4 = f_0 \pm$		0.1	4.5	60	kHz
$f_5 = f_0 \pm$		0.11	5	70	kHz
$f_6 = f_0 \pm$		0.22	10	140	kHz

fig. 7. Required selectivity of the EMI/RFI receiver as a function of frequency range.

The microprocessor-controlled receiver has its own intelligence and combines the proper RF and IF attenuation for optimum dynamic range. It is theoretically possible to increase the IF attenuation rather than the RF attenuation. As in the manually operated receiver, the two functions are not tied together, and the inexperienced operator may not be aware that the intermodulation distortion products can only be reduced by using the RF attenuator.

In order to monitor the actual overload, special detectors are placed after the mixers, because modern receivers use a first IF approximately twice the maximum receiving frequency, the first IF of the test receivers can be expected to be in the vicinity of 70 to 80 MHz. The second IF is then substantially lower (between 9 and 11 MHz), depending upon the receiver, and sometimes even a third IF (30 kHz) for the very narrow bandwidth requirements is used. This design requires two monitoring stages after the mixers to make sure no overload occurs.

time constants

EMI receivers are also distinguished by specific values of detector attack and decay time constants, with typical values being 1 and 160 milliseconds, respectively, in the 0.15-30 MHz frequency range. A good communications receiver uses totally different time constants. In the SSB/CW mode, the attack time would probably vary between 3 and 15 mS, depending upon the manufacturer, and the discharge time constant would be in the vicinity of 200 mS to 10 seconds selectable. The 1 mS attack time for the pulse receiver is too fast and will result in a "quasi-peak" reading, which for the EMI receiver is desirable

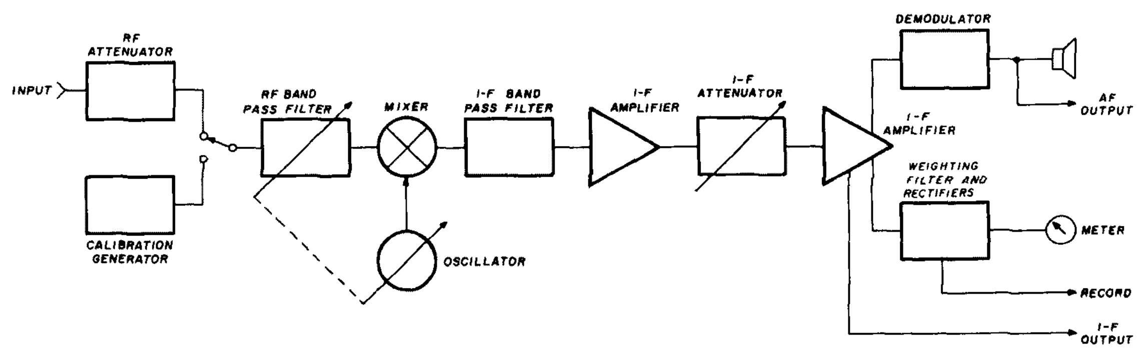


fig. 8. Block diagram of an EMI/RFI receiver.

but would lock up the AGC in a communication receiver each time an unwanted noise spike occurred.

Some manufacturers have chosen to make the EMI/RFI receivers *more universal* by adding built-in detectors for the communication mode as well. The ESH2 and ESH3 have this flexibility.

IF bandwidth

Special IF bandwidths are needed in EMI/RFI receivers; 200 Hz is chosen for the lowest frequency range (10 kHz-150 kHz) 9 kHz for 0.15-30 MHz and 120 kHz for 30-1000 MHz.

At 200 Hz, the bandwidth for the frequency range of 10 kHz to 150 kHz almost requires a triple conversion receiver in order to obtain narrow bandwidth with a good shape factor.

At 30 kHz, the 200 Hz bandwidth filter is more likely a mechanical filter than a crystal filter, as the cost otherwise would be prohibitive.

amplitude calibration

There are two ways to calibrate the receiver. One is to use a pulse generator, such as the ones manufactured by Schwarzbeck, (models IGM 2913, 10 kHz to 30 MHz, and IGU 2912, 25 MHz to 1000 MHz) which operate over a fairly wide pulse rate. With a calibrating pulse of 0.316 microvolts per second and a repetition frequency of 100 Hz, the frequency range of 150 kHz to 30 MHz can be covered. The particular calibration voltage should give a 0 dB reading on the meter.

Sine wave calibration is also possible. This requires a second generator which can be provided inside the instrument. The sine wave output is a good cross-reference for the calibration of the pulse generator.[2] As the calibration of the instrument depends upon these signal sources, it is important that these signal sources be built in such a way that aging effects, temperature, and voltage variations do not affect them. Modern special feedback circuits can solve this problem.

In the case of automated receivers, like Rohde & Schwarz ESH3, the built-in microprocessor, together with the random access memory, allows the development of a scanning program in which the receiver is calibrated over the entire frequency range, and the actual error is stored in memory. As measurements are made, the receiver uses a "look-up" table to add the correction factor. This is convenient because the operator does not have to worry about the accuracy of the receiver.

A manually operated receiver has to be calibrated for each major frequency change, which can be time-consuming since the values also have to be written down for future use. A word of caution: it should be remembered that the frequency synthesizer also is an important factor in receiver performance. The noise sideband of the synthesizer and its inherent spurious performance have to be good enough to prevent any spurious frequencies or sidebands from appearing and giving erroneous readings. Therefore, the reference suppression and all mixing products have to be suppressed sufficiently.

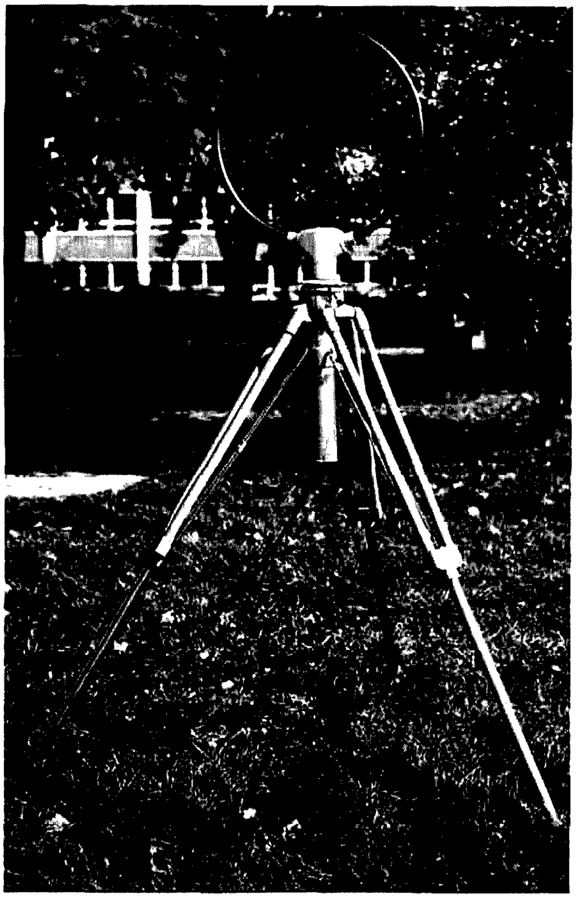


fig. 9. Loop antenna system recommended for EMI/RFI testing.

antennas

The use of tuned antennas is rare at lower frequencies (between 10 kHz and 150 kHz). In this frequency range it is better to use loop antennas or active antennas. Again, it is important to make sure that the dynamic range of the active antennas are sufficient. While the test site has to be properly designed and reflections have to be avoided it should be mentioned here that if an active antenna is used, its dynamic range must be sufficient.

For frequencies above 20 or 30 MHz, reference dipoles or logarithmic periodic antennas may be used, depending upon the particular frequency. It would be best to look up the particular recommendations and

requirements by CISPR and VDE/FTZ. (To measure conducted interference, current probes and absorbing clamps can be connected to the receiver, but this will not be discussed here, since this article is limited to discussion of the receiver itself.)

Fig. 9 shows a loop antenna; **fig. 10** shows an active rod antenna; **fig. 11** shows a log-periodic antenna for VHF/UHF.

conclusion

The EMI/RFI receiver is a more sophisticated and, therefore, more expensive receiver than standard communication receivers. While it is possible to incorporate features to make the reception of communication transmission possible, which is useful for signal identification, then overall accuracy, special pulse response behavior, and the necessary preselector make the receiver more complicated and, thus more expensive. EMI/RFI receivers should be offered in both manual and automated versions to fit varying budgets. However, if large quantities of data must be handled, the automated version is the more logical choice.

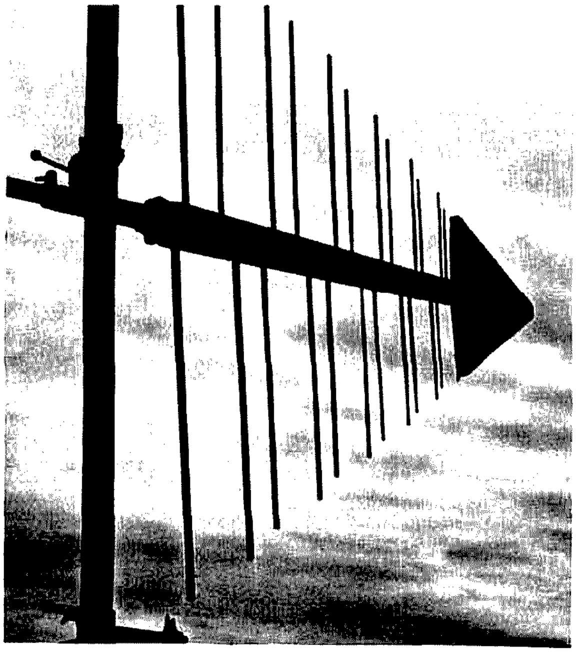


fig. 11. Logarithmic periodic antenna for EMI/RFI testing up to 1000 MHz.

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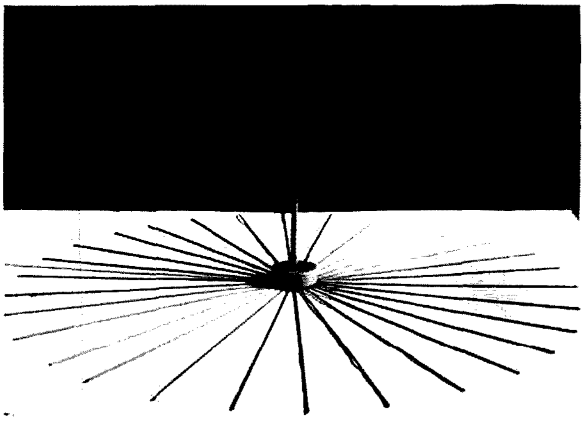

fig. 10. Rod antenna recommended for EMI/RFI testing for frequency range up to 30 MHz.

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ham radio

a simple shortwave broadcast receiver

Most Amateur transceivers just cover the ham bands. Because I wanted something with which I could listen to international shortwave newscasts, I constructed this simple shortwave broadcast receiver that tunes the 6 and 9 MHz SW bands and is inexpensive to build. Building it takes a little skill and you'll need a few test instruments to get it going, but I'm sure you'll enjoy listening to something you've built completely on your own.

To keep construction simple and costs down, I based the circuit on the J.W. Miller 8901-B and 8902-B IF amplifier and detector (mechanical assemblies are illustrated in **figs. 1A** and **1B**). This unit contains three stages of IF amplification and a diode detector inside the capsule. The output is more than enough to drive a 2N2222 audio stage, which in turn drives an LM380N audio chip at several watts output. This in turn drives a 12-inch loudspeaker.

In order to keep the circuit simple and eliminate self-oscillation problems which would require decoupling and shielding, no RF stage was used. The antenna just goes through a double-tuned circuit directly into the 40673 mixer. The large capacity 365-365 gang-tuning condenser eliminates the need for bandswitching for the input stage since it tunes the entire 6 to 9 MHz band. The only drawback is that it has very sharp tuning, and care must be used to make sure the input tuning is on the station you are listening to. With this broad tuning range, it is possible to tune on an image station 455 kHz away. If the panel is marked, there should be no trouble making the tuned circuits track and tune to the same frequency, once the two input slugs are adjusted as described in the following section.

All of the coils in the receiver are wound on National XR-50 coil forms; the number of turns indicated is only approximate, since lead length to the bandswitch can vary. I used a two-position tone control switch for bandchanging.

By Ed Marriner, W6XM, 528 Colima Street, La Jolla, California 92037

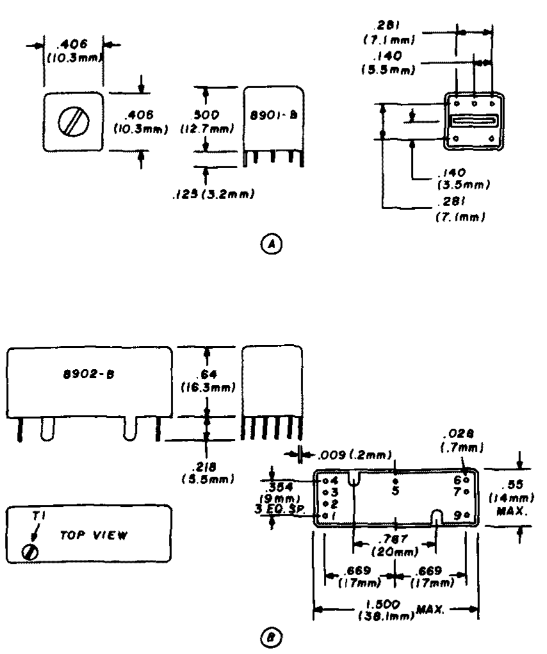


fig. 1. **(A) 8901-B module mechanical assembly; (B) 8902-B module mechanical assembly.**

The receiver was built on a California chassis (No. 122) that measures 4.5 × 8.5 inches (11.43 × 21.59 cm). Using etched boards, everything fits on the top of the chassis, making it easier to work on. Spacers are needed to keep the printed circuit board high enough to prevent the RG-174/U from being punctured by wires projecting from below.

The variable oscillator could probably be made without the emitter follower. However, experience suggests that the isolation between stages is necessary to prevent oscillator pulling. Finally, the oscillator is mechanically tuned using a Jackson Ball 5:1 drive. A schematic of the completed receiver is provided in **fig. 2**.

construction and alignment

The power supply was built first, followed by the audio section. Once I knew the audio section was working I added the IF and mixer stage. The antenna-tuned circuit was last and perhaps the most difficult to adjust. Here a grid dip oscillator is a must. I unsoldered the 7 pF coupling capacitor and tuned each coil separately on 6 MHz for best tracking. The 15 pF trimmer was primarily used to reduce any "mistracking" when different antennas were attached.

table 1. Abbreviated list of stations that transmit in the 6 and 9 MHz bands.

frequency (kHz)	station	frequency (kHz)	station
5955	Voice of Nicaragua	9360	Radio Madrid, Spain
5975	BBC – London	9410	BBC
5985	Radio China	9510	BBC Northern Ireland
6005	BBC	9540	Radio Nederland
6020	Habana, Cuba	9545	Radio Germany
6030	Voice of America	9565	Voice of America
6040	Voice of America	9580	Radio South Africa
6050	HCJB – Quito, Ecuador	9590	Radio Nederland
6060	Radio Haban, Cuba	9630	Radio Spain
6065	Radio Madrid, Spain	9635	Radio Moscow
6070	WYFR ("Family Radio")	9650	Voice of America
6075	WYFR ("Family Radio")	9670	Voice of America
6095	HCJB – Quito, Ecuador	9700	Voice of America
6115	Radio Moscow	9720	HCJB – Quito, Ecuador
6115	Voz de Llanos, Colombia	9735	BBC
6120	BBC	9745	HCJB – Quito, Ecuador
6125	BBC and Voice of America	9755	Voice of America
6140	*Radio Canada*	9780	Radio Moscow
6155	Voice of America	9810	Radio Moscow
6165	Radio Nederland	9825	BBC
6170	BBC	9835	Radio Budapest
6175	BBC	9915	BBC
6195	Radio Canada		

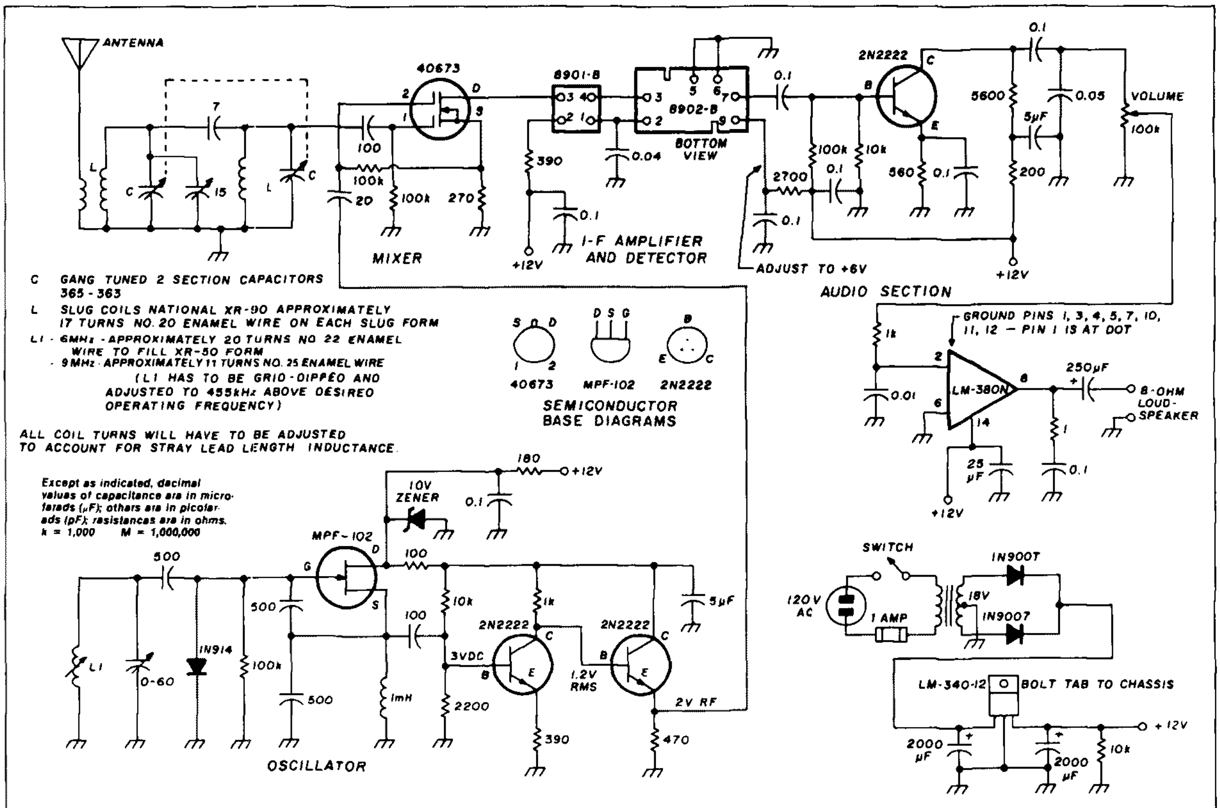


fig. 2. Two shortwave band receiver schematic diagram.

the weekender

(continued)

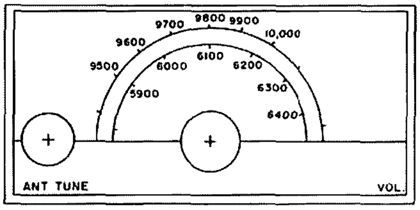


fig. 3. Frequency scale template for the broadcast receiver.

If you have a signal generator modulated by an audio tone it helps to align the 455 kHz IF and approximately calibrate the main frequency dials. If your signal generator is not accurate, then the dial should be checked against another receiver and recalibrated if necessary. (A frequency scale template is provided in **fig. 3**.)

Since only the 6 and 9 MHz bands are used, adjusting the receiver should not be much of a problem. For daytime listening, a coil could be wound for the 11 MHz band.

operation

This receiver is appropriate for language practice and for listening to the news, which most shortwave stations give on the hour. Several religious stations also come in strong: two are WYFR "the Family Radio station," and HCJB in Quito, Ecuador. In the early morning the Japanese stations are particularly strong. I have listed some of the stations active on the two bands and actually heard all those listed in **table 1**.

where to get parts

Be sure to include an SASE when writing for catalogs or information. Transistors and other parts may be obtained from:

Circuit Specialists
Box 3047
Scottsdale, AZ 85257

BCD Radio Parts Co.
P.O. Box 06017
Fort Myers, FL 33906-6017

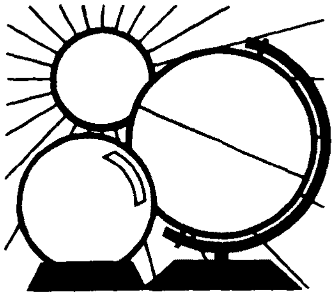
Radio Shack

Knapp Inc.
4750 96th St. N.
St. Petersburg, FL 33708

J.W. Miller/Bell Industries
19070 Reyes Avenue
P.O. Box 5825
Compton, CA 90224

A useful reference for sources of electronic parts is *Radio Electronics Buyers Guide* ($6.95 postpaid), available from Ham Radio's Bookstore, Greenville, New Hampshire 03048.

ham radio



DX FORECASTER

Garth Stonehocker, KØRYW

The proximity of November to the winter solstice, a time when the geomagnetic field is quiet, normally means undisturbed conditions. But because this year is one of several in the sunspot decline, many disturbances can be expected this month. Therefore, we can expect unusual propagation conditions to occur, with the emphasis on the following days: November 7th through 12th, 17th through 20th, and on the 25th and 30th.

vhf-ers take note

Many days of meteor showers will occur between October 26th and November 22nd, with a shower maximum from the 3rd through the 10th at a rate of ten per hour. This shower is known as the *Taurids*. Lunar perigee is on the 1st and 26th, and the full moon on the 20th.

winter season DX

November through February constitutes the winter DX season. Because the D and E regions of the ionosphere receive less energy from the sun in the northern hemisphere during this time, less ionization occurs. Therefore the daytime attenuation of radio signals in winter is lower than during the rest of the year. At the same time, ion production each day is better able to drift and diffuse up into the F region of the ionosphere. The result is an increased range of operating frequencies between the lowest and the maximum usable frequencies (LUF-MUF). The maximum usable frequency rises rapidly as the sun rises each day, peaking just after noontime. The frequency diminishes in the late afternoon, evening, and through the night to a low value just before dawn the next day. The exception to this situation is for locations nearer to the equator, where the F region ionization continues to drift and diffuse up during the afternoon and evening to become the transequatorial maximums described in last month's column. The maximum usable frequency peak reached each day and the depth of the predawn minimum frequency of the next morning are related to the solar flux each day. The higher the flux during the day, the higher the frequency and the lower the dip the next morning.

Wintertime DX provides openings with these characteristics:

- Better daytime signal strengths on the lower frequencies
- Nighttime DX openings earlier each day in the evenings
- More frequent transequatorial paths toward the south
- Higher signal strengths on all bands most of the time.

band-by-band summary

Ten and fifteen meters will be open for F_2 long skip *and transequatorial* one-long-hop propagation. Worldwide DX is prevalent from after sunrise until well after sunset most days, especially during periods of high solar flux conditions and moderate geomagnetic field disturbances.

Twenty meters will be open most days and nearly throughout the night to some areas of the world. This mode follows the sun across the sky: east, south, then west with long skip of 1000-2500 miles.

Thirty meters is a day and night band. The day portion should be similar to 20 meters; signal strengths, however, may decrease during midday on some days of higher solar flux values. This band will also be usable well into the night and often through the night. Problem nights will probably follow high solar flux days and be related to the deep dip of MUF an hour or so before dawn. The distances covered on this band might exceed 80-meter nighttime paths while being less than 20-meter daytime paths.

Forty meters, like 30 meters, is a transition band with all-night propagation as well as some short-skip conditions during the daytime. Most areas of the world can be worked from darkness until just after sunrise. Hops shorten to about 2000 miles on this band, but the number of hops can increase since the signal attenuation is low at night.

Eighty meters, traditionally the ragchewing band, is also good for distant operation. The band operates much like 40 meters, except in that the hop distances shorten to around 1500 miles at night and even less during the daytime. Because the noise is so low, this band is a pleasure to work during this time of year. The path direction follows the darkness across the earth — east, south, then west. Lots of QRM can be expected, however. (*Remember, the DX window* is 3790-3800 kHz.)

One-sixty meters will be similar to 80 meters, with skip hops reduced to

NOVEMBER

WESTERN USA

GMT	PST	N (ASIA FAR EAST)	NE (EUROPE)	E (S. AFRICA)	SE (S. AMERICA)	S (ANTARCTICA)	SW (NEW ZEALAND)	W (OCEANIA AUSTRALIA)	NW (JAPAN)
0000	4:00	20	20	15	10	15	10	10	20
0100	5:00	20	20	15	10	15	10	10	20*
0200	6:00	20	20	15	10	15	10	10	15
0300	7:00	20	30	20	10	15	10	10	15
0400	8:00	20	30	20	15	15	10	10	20
0500	9:00	*20*	*30*	20	15	15	10	15	20
0600	10:00	*20*	*30*	20	15	20	15	15	20
0700	11:00	*20*	*30*	20	20*	20	15	15	20
0800	12:00	*20*	*30*	20	*20*	20	15	20	30
0900	1:00	*20*	*30*	*20*	*20*	*20*	20	20	30
1000	2:00	*30*	*30*	*20*	*20*	*20*	20	20	*30*
1100	3:00	*30*	*30*	*20*	*20*	*20*	20	20	*30*
1200	4:00	*30*	*30*	*20*	*20*	*20*	20	20	*30*
1300	5:00	*30*	*30*	*20*	*20*	*20*	20	*20*	*30*
1400	6:00	*30*	*20*	*20*	*20*	*20*	*20*	*20*	*30*
1500	7:00	*30*	*20*	15	*20*	*20*	*20*	*20*	*30*
1600	8:00	*30*	*20*	15	20	*20*	15	*20*	*30*
1700	9:00	30	20	15	15	*20*	15	15	*30*
1800	10:00	30	20	10	15	15	10	15	*20*
1900	11:00	30	20	10	15	15	10	15	20
2000	12:00	30	15	10	15	15	10	10	20
2100	1:00	30	15	10	10	15	10	10	20
2200	2:00	20	15	10	10	15	10	10	20
2300	3:00	20	20	15	10	15	10	10	20

MID USA

MST	N (ASIA FAR EAST)	NE (EUROPE)	E (S. AFRICA)	SE (S. AMERICA)	S (ANTARCTICA)	SW (NEW ZEALAND)	W (OCEANIA AUSTRALIA)	NW (JAPAN)	CST
5:00	30	20	15	10	15	10	10	20	6:00
6:00	30	20	15	10	15	10	10	20	7:00
7:00	30	30	15	10	15	10	10	20*	8:00
8:00	30	30	20	10	15	10	10	15	9:00
9:00	40*	30	20	15	15	15	15	15	10:00
10:00	40	30	20	15	15	15	15	20	11:00
11:00	40	*30*	20	15	20	15	15	20	12:00
12:00	40	*30*	20	20	20	15	20	20	1:00
1:00	*30*	*30*	20	20	*20*	20	20	*30*	2:00
2:00	*30*	*30*	20	20	*20*	20	20	*30*	3:00
3:00	*20*	*30*	20	20	*20*	20	*20*	*30*	4:00
4:00	*20*	*30*	*20*	*20*	*20*	20	*20*	*30*	5:00
5:00	*20*	*30*	*20*	*20*	*20*	*20*	*20*	*30*	6:00
6:00	*20*	*20*	*15*	*20*	*20*	*20*	*20*	*30*	7:00
7:00	20	*20*	15	*20*	*20*	*20*	*20*	*30*	8:00
8:00	20	*20*	15	20	*20*	*20*	*20*	*30*	9:00
9:00	20	20	15	15	*15*	15	*20*	*30*	10:00
10:00	20	20*	10	15	*15*	15	15	*30*	11:00
11:00	20	20*	10	15	15	15	15	*20*	12:00
12:00	30	20	10	10	15	15	15	*20*	1:00
1:00	20	20	10	10	15	10	10	20	2:00
2:00	20	20	10	10	15	10	10	20	3:00
3:00	20	20	10	10	15	10	10	20	4:00
4:00	30	20	15	10	15	10	10	20	5:00

EASTERN USA

EST	N (ASIA FAR EAST)	NE (EUROPE)	E (S. AFRICA)	SE (CARIBBEAN S. AMERICA)	S (ANTARCTICA)	SW (NEW ZEALAND)	W (OCEANIA AUSTRALIA)	NW (JAPAN)
7:00	30	20	15	10	15	10	10	20
8:00	30	20	15	10	15	10	10	15
9:00	30	30	15	10	15	10	15*	15
10:00	30	30	20	10	15	15	15	20
11:00	40	30	20	15	15	15	15	20
12:00	40	30	20	15	15	15	20	20
1:00	*40*	30	20	15	20	20*	20	30*
2:00	*40*	*30*	20	15	20	20	20	*30*
3:00	*30*	*40*	20	20	*20*	20	*20*	*30*
4:00	*30*	*40*	20	20	*20*	20	*20*	*30*
5:00	*20*	*40*	20	20	*20*	*20*	*20*	*30*
6:00	*20*	*30*	*20*	*20*	*20*	*20*	*20*	*30*
7:00	*20*	*30*	*20*	*20*	*20*	*20*	*20*	*30*
8:00	*20*	*30*	*15*	*20*	*20*	*20*	*20*	*30*
9:00	20	20	15	20	*20*	*20*	*20*	*30*
10:00	20	20*	15	15	*20*	*20*	*20*	*30*
11:00	20	20*	10	15	*20*	*20*	*20*	*30*
12:00	20	20*	10	15	15	15	*20*	*30*
1:00	20	20	10	15	15	15	15	*30*
2:00	*20*	20	10	10	15	15	15	*20*
3:00	*20*	20	10	10	15	10	15	20
4:00	30	20	10	10	15	10	10	20
5:00	30	20	10	10	15	10	10	20
6:00	30	20	15	10	15	10	10	20

The *italicized* numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during 'normal' hours.

*Look at next **higher** band for possible openings.

about 1000 miles. It will provide good DX for late night and early morning DXers. Stations in many areas of the country can now run higher power. (Once again, please keep the DX windows — 1825-1830 kHz and 1850-1855 kHz — free of local contacts.)

last-minute forecast

During November expect the higher frequency bands — 10 through 30 meters — to be best during the middle of the month; a solar radio flux maximum is expected during that time. Maximum flux is also possible on the 25th which would mean a good, long DX holiday weekend. (Monitor radio station WWV for geophysical data at 18 minutes after the hour on 2.5, 5, 10, 15 or 20 MHz to update this forecast.) The lower frequency bands are expected to be good throughout the month, but somewhat better during the predawn hours at the beginning and end of the month.

ham radio

short circuits

building blocks

In KB0CY's article, "Audio Filter Building Blocks", (July, 1983), **fig. 2** (page 76) should show all +12V connections going to pin 7 of the four LF356's. One LF356 (upper center) shows pin 1 as the +12V connection; this is incorrect.

briefcase bobtail

In the directions for a "Briefcase Bobtail" given by Paul M. Rich (Comments, July, 1983, page 12), the twelve turns of No. 14 wire should be spaced 1/4 inch apart, not one inch apart. The call sign HH2KR was incorrectly given as HH2DR.

Bobtail curtain

In part one of the Bobtail curtain series by W6BCX (February, 1983, page 82), an article on the Bobtail is referred to in the April, 1948, issue of *CQ*. The correct date of publication is March, 1948.

technical forum

Have a technical question? Ask your fellow Amateurs through Technical Forum. You'll reach knowledgeable readers ready to help you find the answer you're looking for. *Have an answer?* Send it in. Every month we'll award a specially selected book from Ham Radio's Bookstore to the writer of the best response to a question previously posed in this column. Address your questions and answers (typewritten, if possible) to Technical Forum, *ham radio magazine*, Greenville, New Hampshire 03048.

bunny hunt

In the May, 1983, Technical Forum, N3BEK raised the problem of RFI on 160 meters from a local broadcast station.

May I suggest he try using the station's field intensity meter to go on a low-frequency "bunny hunt." It is not unusual for the down guys on utility poles to unintentionally provide 60 Hz rectification at corroded joints and create fluorescent light-type noise in high-frequency receivers. Utility wires, down guys, grounds from pole-mounted transformers, and the like can all have partially corroded splices and connections that are good at power line voltages but act as semi-conductors when exposed to the 1-10 volts of RF (field intensity in V/M). This type of re-radiation is common in many AM antenna systems, and has to be tuned out at the source of re-radiation; that particular parasitic element has to be made non-resonant at the carrier frequency in order to make the antenna system of the station produce the desired radiation pattern.

It is entirely possible that a downguy, power drop messenger cable, or the ham antenna might have a discontinuity that would produce the "mystery station on 160 meters." The temperature/frequency relationship mentioned in the column would indicate a change in the resonant frequency as the element gets warm; it gets longer due to expansion, and the self-resonance lowers. The fact that the station uses asymmetrical modulation has no bearing on the situation other than developing 3 dB additional sideband power on modulation tips. Assuming mixing is taking place by the diode-type action mentioned above, at a non-linear portion of that diode, and the self-resonance of the antenna producing the carrier and higher sideband energy, this would account for the unintelligible audio mentioned in the letter. The cure can be as simple as cleaning the connection or adding a resonant circuit to detune it from the 160 meter band or 1500 kHz station. Judging by the report of the station's engineering people and the intensity at N3BEK's QTH, my guess is that it's close to the shack. But, since you have to find the source to cure it, that's where the field intensity meter is helpful. It should be able to detect signals to fractions of a microvolt. Hope this helps. Good luck. — **Ed Karl, KØKL**

mysterious spur solved

In response to my letter to Technical Forum (May, 1983), I received a telephone call from Robert Schantz, a Los Angeles broadcast radio consultant. After determining that the local 50 kW transmitter was manufactured by Continental, he said he was familiar with the design of that particular asymmetrically-modulated transmitter. He has experienced the generation of spurs from transmitters of that design — in one case, 250 kHz in the broadcast band. He said the Continental transmitter includes a complex feedback circuit in the modulator that is critical to adjustment. When the adjustment allows the generation of a spur, the frequency of that spur depends on the characteristics of the antenna system.

It is his conjecture that small changes in antenna characteristics, along with changes in temperature, are causing the frequency of the spur to shift.

My money is on Mr. Schantz's solution. — **Jack Geist, N3BEK**

too many turns

Like WB2NTQ (June, 1983), I have also tried to create high impedance (above 100 ohms) transformers with powdered iron toroids and failed. Here is why.

The problem in WB2NTQ's transformer is in his secondary winding, which at 29 MHz, is operating above its natural resonant frequency. The usable bandwidth for a transformer on the high end is determined by the self resonance of the windings. This resonance occurs when a winding is effectively 1/4-wavelength long. Due to the dielectric constants of wire insulation and core material, and the capacity between adjacent turns the total wire length in a resonant winding can be significantly less than 1/4 wavelength in air.

The low frequency limit of a transformer is determined by the inductive reactance of the transformer winding becoming lower than five or so times the source or load impedance. Small cores of high permeability material produce transformers with the greatest bandwidth.

The largest core I can find data on is a T200 size. Each turn is 1.85 inches on that core and could be longer on a larger toroid. Total length on WB2NTQ's high impedance winding is then about 135 inches (ignoring wire size and winding looseness). This is a quarter wave at 21.86 MHz, neglecting dielectric effects. Thus the capacitive reactive term is in parallel with the resistive component on the input side.

There are just too many turns on the transformer. The low permeability of the powdered iron core makes the design all the more difficult. I'd first cut the secondary length down to about 1/8-wavelength at the maximum desired frequency. That would

be about 49 inches and would wind twenty-six turns. I would wind these as thirteen turns bifilar connected series aiding. The center tap would be the ground side of the 50 ohm connection. This would be a simple 75 to 300 ohm transformer. To get 50 ohms I would tap one side of the bifilar winding at 10 or 11 turns out from the center tap. Ten turns would give a turns ratio of 10 to 26 or an impedance ratio of 50 to 338 ohms. Tapping at 11 turns would give a ratio of 50 to 279 ohms.

The low frequency end of this transformer would be at the frequency where the inductance of the 11-turn section has an inductive reactance of 5 × 50 = 250 ohms. On a T200-2 this winding inductance is about 1.3 microhenries. To the limit of the accuracy with which I can read my Shure reactance rule, this happens at 30 MHz!

A larger core would give more inductance per turn but at the same time would have more wire in a turn. High impedance, large size and low permeability together prevent this core from performing adequately. The powdered iron toroid just is not appropriate for such a high impedance winding, but might function if constructed within these limits for a single band.

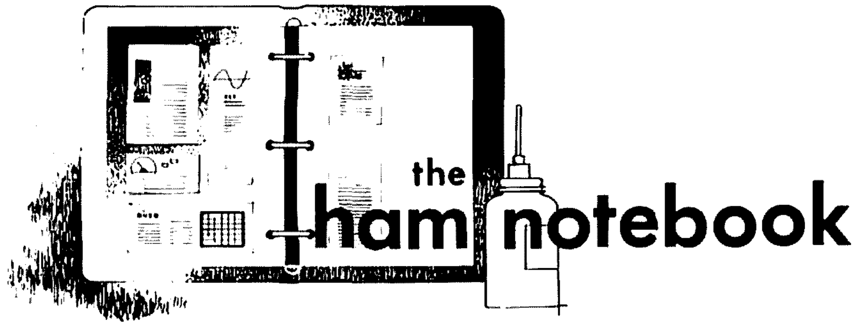
A better core for this application might be either an F-240 or F-114 in Q1 (mix 61) or Q2 (mix 62) ferrite. The smaller core in Q1 would require five turns on the 50 ohm section and twelve turns total on the 300 ohm winding for a minimum operating frequency of 14 MHz. I would wind this as six bifilar turns connected series aiding with the primary tap at five turns. The windings should be spread uniformly around the whole core. The impedance ratio would be 50 to 288 ohms. On this core a winding would have a length of about one inch per turn. Twelve secondary turns would be about 12 inches long and should work well beyond 50 MHz.

To achieve wide bandwidth it is necessary to use a high permeability core material to extend to low frequency end of the pass band with a minimum winding conductor length. At the high frequency limit the core is practically uncoupled from the winding and only the core's dielectric constant is significant. The dielectric constant for ferrite can be high, so isolating the winding from the core can help extend the high frequency end as long as the winding conductor length does not grow too much at the same time. Thick wire insulation reduces the inductive coupling between adjacent bifilar turns at the high frequency end of the pass band. A better winding would be made of enameled wires twisted together for the bifilar winding and then covered with a heavy walled Teflon™ insulating tubing. — **Gerald A. Johnson, K0CQ**

standing-wave indicator

I have just bought a standing-wave indicator, type B812A, manufactured by FXR, Inc., of Woodside, New York. The unit is not functioning at present and I wonder whether anyone might have a circuit diagram or other information which would help me get the unit working.

After reading the interesting article on this type of unit by Bob Stein, W6NBI, in the January, 1977, issue of *ham radio*, I feel the device would be a very useful addition to my workshop. — **Arthur Williams, GW8FKB**



improved stability and dial calibration for the Heathkit HW-8

The HW-8 transceiver exhibits approximately 1500 Hz drift in transmit and receive frequency when the supply voltage varies over a range of 10 to 13.5 VDC. This results in CW chirp when using a poorly regulated supply, such as a weak, dry battery. Additionally, even with a well-regulated supply, the VFO dial calibration is in error on all but the 7 MHz band.

Most of the drift and chirp problem is caused by the Heterodyne Oscillator (Q6). The reverse-biased switching diodes in the tuned circuits of all but the selected band exhibit a capacitance which varies with supply voltage. This capacitance, essentially in parallel with the selected crystal, causes pulling of the oscillator frequency. The solution is to regulate the supply voltage to Q6. The small amount of shift which still remains after Q6 is stabilized is caused by the inability of the Zener diode (ZD-1) to fully stabilize the voltage for the Variable Frequency Oscillator (Q2). This can be corrected by replacing the Zener-diode regulator circuit with a Motorola MC7808CP three-terminal regulator integrated circuit.

The VFO dial calibration problem is a matter of fine tuning the VFO and HFO in accordance with the procedure described here. The Heathkil procedure does not calibrate the frequency of the HFO; it also does not switch in the offset capacitor (C55) during VFO calibration so that the dial will read transmit frequency.

modification procedure

Remove the following resistors: R78, R81, R82, R84, R85, R87, R88, and R91 (see fig. 1).

Install 7.5-volt, 1-watt (SK-3059) or equivalent, *Zener diodes (anode lead to ground)* in the positions formerly occupied by R81, R84, R87, and R91 (100k resistors).

Install 470 ohm, 1/4-watt resistors in the positions formerly occupied by R78, R82, R85 and R88 (1k resistors).

Install a 0.01 μF, 25 VDC ceramic capacitor on the foil side of the main PC board. Solder one lead to the junction of R36 and the yellow wire which attaches to point *B*. Solder the other lead of the capacitor to a nearby ground foil.

Remove ZD-1 and R33 (470 ohm). Drill a 1/32 inch hole midway between the two holes from which R33 was removed. Install the MC7808CP voltage regulator as follows:

Input *B* lead to R33 hole which ties to 13.4-volt line; insert common *C* lead through the drilled hole and output *E* lead to R33 hole which ties to C52 and R3 (47 ohm).

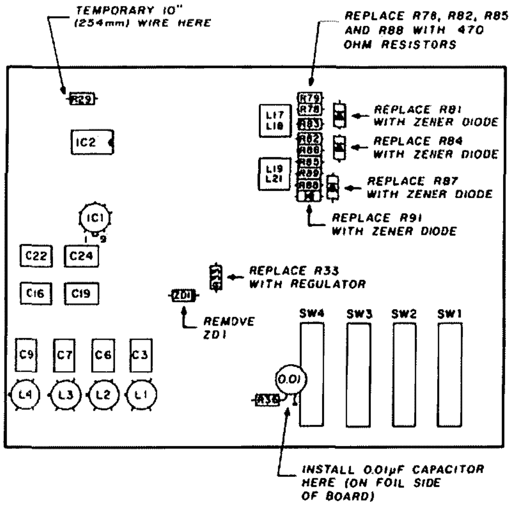


fig. 1. Component drawing indicates modifications to be made.

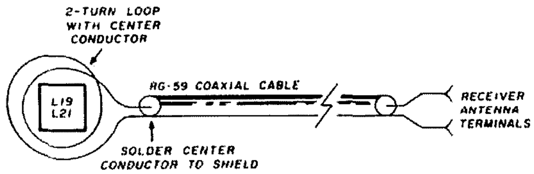


fig. 2. 2-turn pick up loop is used for fine alignment.

Solder and clip the excess from the *B* and *E* leads. Slip a piece of insulation over the *C* lead and solder the lead to a nearby ground foil. Be sure that it does not short to other foil leads.

fine alignment procedure

Make a pick-up loop as shown in fig. 2 and place it around L19/21. Connect the opposite end to the antenna terminals of a calibrated receiver capable of tuning 12 to 30 MHz.*

Press the 3.5 MHz bandswitch.

Tune the calibrated receiver to 12.395 MHz.

Adjust L17 (bottom slug) for zero-beat.

Press the 7.0 MHz bandswitch.

Tune the calibrated receiver to 15.895 MHz.

Adjust L18 (top slug) for zero-beat.

Press the 14.0 MHz bandswitch.

Tune the calibrated receiver to 22.895 MHz.

Adjust L19 (bottom slug) for zero-beat.

Press the 21.0 MHz bandswitch.

Tune the calibrated receiver to 29.895 MHz.

Adjust L21 (top slug) for zero-beat.

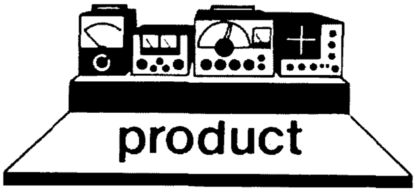
Temporarily attach a 10-inch piece of wire to the end of R29 (22k) which connects to point WW. Connect the other end of the wire to one of the ON/OFF switch terminals. This will cause the antenna relay to close and the receiver to mute.

Realign the VFO as described in the Heathkit instruction manual, page 62.

Remove the temporary wire and reinstall the cabinet cover. This completes the modification and alignment.

Robert W. Lewis, W3HVK

*If a calibrated receiver for this frequency range is not available, a frequency counter can be used. The output of the heterodyne oscillator can be picked off at the emitter of Q7, preferably through a 0.001 to 0.01 μF coupling capacitor. The pick-up loop described for use with a receiver likely will not provide enough signal to drive a frequency counter. — Editor



REVIEW

Soar model 5025 digital multimeter

A number of different multimeters have crossed my desk in recent months and I must admit it has been fascinating to see and use the latest state-of-the-art equipment. The newest unit on my desk for review, however, is quite different from the others.

The Soar Model 5025 is not just a nuts-and-bolts measuring device. It incorporates a unique comparator circuit that can be used in a number of different ways.

One of the greatest advantages of the "new breed" of multimeters is extensive use of specially designed chips. The Soar 5025 has a unique 80 pin LSI chip that keeps overall parts count down while ensuring long term stability and accuracy. Only by examining the schematic can one fully appreciate how LSI has changed the complexion of equipment design and utilization.

general specifications

The 5025 has an easy-to-read, low current consumption, LCD readout with a maximum reading of 1999. The readout also has annunciators to audibly alert for function, unit polarity, decimal, low battery, continuity and diode test. The unit is mounted in a rugged ABS plastic case with a U-bracket handle/tilt stand and is fully shielded from RFI/EMI. The probes are designed for safety to reduce the chance of an accidental shock when being used. One of the "neatest" innovations of this new breed of multimeters is the automatic ranging feature. When doing a number of different voltage or resistance readings, this feature is quite a time saver.

Battery life is estimated at >300 hours with alkaline batteries and >200 hours with regular zinc carbon batteries. Four size "C" batteries or a portable AC adapter may be used. The model 5025 also incorporates a built-in overload protection for all ranges with surge protection up to 6000 volts.

comparator circuit

Besides standard ohms, volts, and current measurements, the Soar Model 5025 also has a built-in comparator circuit. The comparator can be used on all measurement ranges and was designed with production and QC testing in mind.

To use the comparator circuit, you select the designed high and low figure on the thumbwheel switches above the LCD readout, punch in the proper range to be tested, and press the "compare" switch. The 5025 will then measure the parameter in question. If the value being measured is within the limit set, the beeper will sound and the value will appear in the LCD readout. If the value is either above or below the preset limits, the beeper will not sound.

use

I had occasion to use the Soar 5025 while troubleshooting a broken radio. I found the U-bracket to be invaluable in getting the multimeter into a position that was easy to see. I also found the 1/2 inch LCD readouts to be a nice feature. I also used it outside to make a number of continuity checks on a vertical antenna ground system. The LCD readouts are easy to read in the sun and the handy beeper assured that there was circuit continuity without the need of looking at the unit.

specifications

Size: 7.25" × 2.25" × 7.125" (186 × 57 × 180 mm)

Weight: 1.9 pounds (850 g) less batteries

Cp. Temp 0°C to 40°C

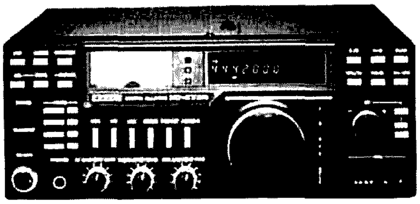
Accuracy guaranteed for 1 year
- DCV ±.25%
- ACV ±.5%
- Resistors 200-200 kilohms ±.25; 2000 K ±1%; 20 M ±2%
- DCA 200μ to 200 mA; 10A ±1.0%
- ACA 200μ – 20 mA ±1%; 200 mA – 10 A ±1.2%

For more information contact NA Soar, 1126 Cornell Avenue, Cherry Hill, New Jersey 08002. RS#313

N1ACH

ICOM transceiver

ICOM's new IC-471A is a 20 MHz coverage base station transceiver for 430-450 MHz. It



NEW products

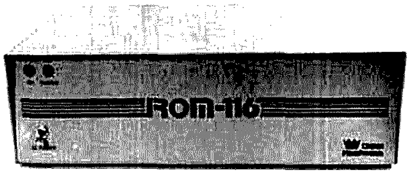
features 10 watt output, 32 full function memories, and built-in sub-audible tone selectable from the main tuning dial.

ICOM system accessories work with the IC-471A. This unit features ICOM's new two-color display, showing frequency digits in white and control functions in red, for easy visibility under all lighting conditions. With the IC-471A, it is possible to scan all frequencies, memories, or modes.

For information, contact ICOM America, Inc., 2112-116th Ave., NE, Bellevue, Washington 98004. RS#311

RTTY/CW interface

The ROM-116 is now distributed and sold exclusively by the Flesher Corporation. The ROM-116 interfaces the Radio Shack TRS-80* models I, III, and IV, and comes with features that include two serial ports, fourteen buffers, split-screen (formatted or unformatted) vertically displayed status, automatic CW/ID, PTT control, Sel-Cal, error correction, text editor, quick break with word mode, word wrapping, preload, two independent callsign buffers (one for user and one for station called), repeat transmission, right hand justification on transmit from the main buffer, time and date with automatic update, and adjustable line length. It also will support a computer printer for hard copy, receive and send RTTY at all standard Baudot and ASCII rates up to 1200 baud; it is TTL compatible, will receive and transmit CW with full break-in mode, CW preload (cassette or disk versions), and cassette or disk save messages or pictures. Several software packages, such as a MAILBOX program (1.4MBO or 3.4MBO) and LOAD HEX (for receiving/



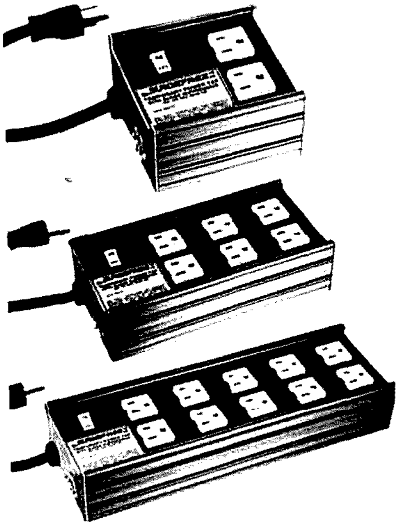
sending disk files on RTTY), are also available. Two versions are marketed; prices will range from $225.00 for the older units to $325.00 for the newer units.

For information, contact Flesher Corporation, at P.O. Box 976, Topeka, Kansas 66601. RS#309

*A trademark of the Tandy Corp.

power center

Ultima Electronics announces the immediate availability of a new state-of-the-art electronic outlet power center designed for fail-safe industrial, residential, and commercial use.



Designated "Surgefree," the new unit features all solid-state electronic circuitry. Compact in size, it can be plugged into any 120 VAC outlet to instantly sense and suppress destructive effects of high-voltage transient spikes and surges to sensitive electronic equipment. It is rated at 15 amps (1875 watt), 125 VAC, with a resettable circuit breaker that protects against accidental power overloads.

Model SF-200, with two sockets, sells for $69.95; Model SF-600, with six sockets, $89.95; Model SF-1000, with ten sockets, $99.95.

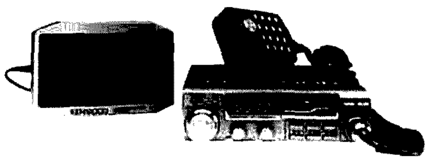
For further information, contact Ultima Electronics Ltd., 59-7 Central Avenue, Farmingdale, New York. RS#308

compact mobile transceivers

Trio-Kenwood Communications has announced the addition of two new ultra-compact models to their line of mobile transceivers.

The 2-meter version, model TM-201A, incorporates microprocessor-controlled operating features in a new lightweight slim-line design. Features include 25 watts of RF output, dual-digit VFO's, five memories, priority alert scan, memory and band scans, lithium battery memory back-up (estimated 5-year life), high-visibility yellow LED display, external speaker, and a 16-key autopatch UP/DOWN microphone. An audible "beeper" confirms operation of selected functions. An optional FC-10 frequency

NEW products



controller allows remote control of frequency, VFO selection, memory recall, and memory channel selection. An optional TU-3 two-frequency tone encoder permits operation on repeaters having different sub-audible tones.

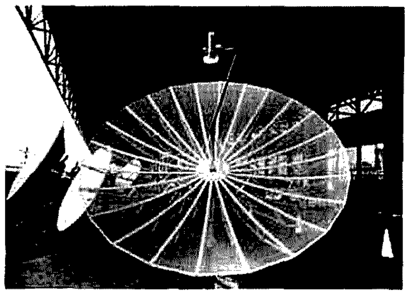
A 70 cm version, the TM-401A, is similar to the TM-201A in features and appearance, and is available with a maximum of 12 watts RF output.

For additional information, contact Trio-Kenwood Communications, 1111 West Walnut Street, Compton, California 90220.

modular dish antenna

The new modular aluminum X-11 satellite TV antenna developed by KLM Electronics, Inc., was named one of the most innovative products for 1983 in the Design & Engineering Exhibition of the Summer Consumer Electronics Show at Chicago's McCormick Place.

The X-11 is the first satellite TV antenna to be selected for recognition at the Exhibition. Because of its size, it was displayed outside the exhibition hall.



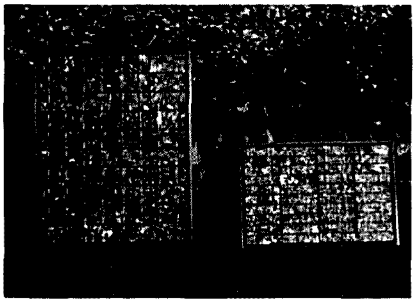
The X-11 is designed to reduce dealer setup time by 60 to 70 percent and to permit easy assembly by the consumer. It handles 100 mph winds and can be shipped by United Parcel Service.

For more information, contact KLM Electronics, Inc., 16890 Church Street, Morgan Hill, California 95037. RS#307

panel discussion

Two new photovoltaic panels, the SX-10 and SX-20, are available from ENCON. Rated at 10 watts; the SX-10 features different current/voltage selections (8 VDC at 1.05 amperes, 13.3 VDC at 0.52 amperes) the ham can wire himself. The SX-20, rated at 20 watts, offers a choice of 8.6 VDC at 2.09 amperes and 17.3 VDC at 1.05 amperes.

The SX-10 and SX-20 can be used for mobile QRP operations or can be permanently mounted for charging batteries. Their life expectancy is 30 years or more, and they're said to be able to withstand a wind load of over 160 MPH and golfball-size hailstones. They're both water- and moisture-proof.



For complete information, contact ENCON, 27600 Schoolcraft Road, Livonia, Michigan 48150. RS#306

precision tips

Six new precision soldering iron tips have been introduced for use with the recently introduced Ungar Series 9000 soldering iron and systems. The new tips are 1/16 and 0.090-inch spade, 1/32 and 3/64-inch screwdriver, 1/32-inch short spade and the 0.020 conical. The list price of each tip is $3.75. All are interchangeable with the modular Ungar System 9300 and System 9000 and 9100 variable-temperature systems.

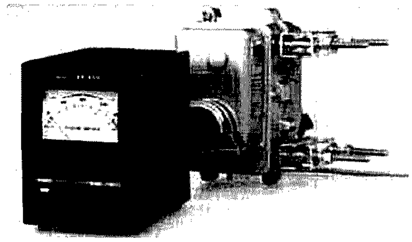
For further information, contact Ungar, P.O. Box 6005, Compton, California 90220. RS#305

elevation rotators

The entire Kenpro product line is again available to Amateurs in the United States. Distributed to local dealers by Spectrum West, the Kenpro line includes the KR 500 elevation rotor, said to be the only dedicated elevation rotator available to retail consumers. The line also includes the KR 2000 RC, described as the strongest azimuth rotator available, with over 10,000 kg/cm torque and the ability to hold over 30 square feet of wind load in a tower configuration.

For more modest applications, Kenpro

NEW products



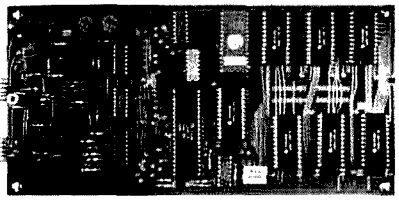
makes the smaller KR 400 and KR 600, which measure 12 and 14 square feet, respectively. A full line of accessories is also available.

The KR 2000 RC lists for $495.95; the KR 400, $149.95; the KR 600, $259.95.

For further information, contact Spectrum West, 5717 N.E. 56th Street, Seattle, Washington 98105. RS#304

packet radio TNC

The Model PK1 TNC from GLB Electronics has a self-contained MODEM and requires only a 12-volt power supply, a data terminal and a radio transceiver for packet operation. The data terminal can be a personal computer, a



"dumb" terminal (keyboard and display), or even a mechanical teletype machine. The terminal interface is RS-232 compatible and self-adapts to ASCII or Baudot and at data rates ranging from 45 to 9600 baud. An adaptor is available for converting mechanical teletype machines to the RS-232 interface. Standard Bell 202 tones are used, with a data rate of 1200 baud, making it compatible with both Vancouver (VADC) and Tucson systems.

Utilizing a Z80A microprocessor, the Model PK1 has 8K of ROM and 4K of RAM as standard equipment. RAM can be readily expanded to 14K, using 2K "byte-wide" memory chips and to 56K via modification using 8K chips. The VADC protocol is available now, and AX.25 is to be released by the end of the year. Conversion to AX.25 is accomplished by means of *exchanging ROM's at nominal cost.*

The Model PK1 is a printed-circuit assembly, measuring 4.5 × 9.4 inches. It's priced at

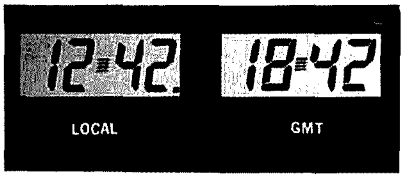
NEW products

$149.95, wired and tested, with documentation and instructions. Power required is a single +12 volt supply at 1/4 amperes. Connecting cables are also available. Documentation is available for $5, refundable on purchase.

For complete information, contact GLB Electronics, 1952 Clinton Street, Buffalo, New York 14206. RS#310

ham clock

The BHC Big Ham Clock features two large 5/8 inch tall LCD modules, one for local time (12 or 24 hour type) and one for GMT. Each clock module can be programmed for the desired combination of month/day, hours/minutes, seconds, *and set to* WWV.

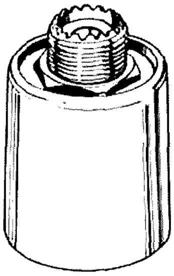


Each module will run from one to three years on the replaceable battery; the two modules are mounted in a black anodized desk top frame. The price is $29.95, plus $1.50 for shipping.

For more information, contact BHC, Inc., 1716 Woodhead, Houston, Texas 77019. RS#303

test mount adapter

The new Larsen Electronics test mount adapter simplifies tracing problems in antenna or radio. It can be used to check the antenna feedline VSWR and the radio power output simultaneously by simply screwing the adapter onto the mount and applying a dummy load. Other uses include use as a coax extension.



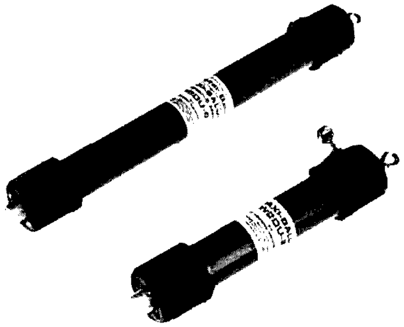
For further information, contact Larsen Electronics, P.O. Box 1799, Vancouver, Washington 98668. RS#312

NEW products

"maxi" baluns

The Unadilla/Reyco/Inline W2DU-(6) and W2DU-(2) baluns handle 3.5 kW power. Model W2DU-(6) is used for 160-6 meter applications while the W2DU-(2) handles those in the 6-1¼ meter range. Pull-apart tensile strength is rated at more than 600 pounds.

The baluns are adaptable to dipoles, inverted-vee, quads and Yagi antennas. Both are contained in weatherproof housing and have built-in lightning arrestors.



The price of each is $24.75. For more information, contact Microwave Filter Co., Inc., 6743 Kinne Street, East Syracuse, New York 13057. RS#302

easy-to-operate shortwave radio

The new Bearcat DX-1000 features direct access keyboard tuning and covers the 10 kHz to 30 MHz spectrum. Ten memory channels allow storage of favorite frequencies for instant recall or for faster "band scanning" during important openings. The digital display measures frequencies to 1 kHz and doubles as a two-time zone, 24-hour digital quartz clock. A built-in timer can be programmed to activate peripheral equipment. Other important features include independent selectivity selection, with 12, 6, and 2.7 kHz IF filters to help separate high-powered stations on adjacent frequencies, plus a two-position noise blanking system that stops Russian interference. It also features an RF attenuator, FM squelch control, tone control, battery backup system, LED indicators for modes and functions, front mounted speaker, "fast" and "slow" AGC, and separate push buttons for selecting AM, LSB, USB, CW, or FM modes. The unit comes with a built-in telescoping antenna for portable use, and includes an SO-239 antenna connector for 50 ohm leads, and a screw connector for "High Z."

The suggested retail price is $599.95. For further information, contact Electra Company, 300 East County Line Road, Cumberland, Indiana 46229. RS#301

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ROHN TOWERS: Wholesale direct to users. All products available. Write or call for price list. Also we are wholesale distributors for Antenna Specialists, Regency, Hy-Gain and Wilson. Hill Radio, P.O. Box 1405, 2503 G.E. Road, Bloomington, IL 61701-0887. (309) 663-2141.

CABLE TV EQUIPMENT, Channel 3 or 4 notch filter $20.00. Information $1.00. Goldcoast, P.O. Box 63/6025G, Margate, FL 33063. (305) 752-9202.

NEED TO CONTACT James Navarchi concerning Yaesu gear. C.T. Huth, 146 Schonhardt St., Tiffin, OH 44883.

CHASSIS and cabinet kits. SASE K3IWK.

TRADE general radio frequency and deviation meter model 1124-A for set of machinist tools from retired machinist, toolmaker. Sell Drake R4-C receiver with filters only $285.00. Collection of over 35 years boxed receiving tubes most at one dollar each. Send your list of needs for availability. W5QJT, 314 N. Resler Dr., El Paso, TX 79912, Apt. One. Tel: 915-581-2017.

VLF-LF preamps, coupler, Loran-C boards. SASE. Burhans Electronics, 161 Grosvenor St., Athens, Ohio 45701.

ELECTRON TUBES: Receiving, transmitting, microwave . . . all types available. Large stock. Next day delivery most cases. Daily Electronics, 14126 Willow Lane, Westminster, CA 92683. (714) 894-1368.

WANTED: Kenwood TS-700SP transceiver and MARS-7600 adapter. C.T. Huth, 146 Schonhardt, Tiffin, OH 44883.

HEARD THE LATEST? You will if you read The Westlink Report (formerly HR Report). Only $22.50 for 26 issues. Fewer words — more tacts. Send $1 for two sample issues (refundable for $3 if you subscribe). Poco Press, Offer #811, 11119 Allegheny St., Sun Valley, CA 91352.

PRE-1946 TELEVISION SETS wanted for substantial cash. Finder's fee paid for leads. Also interested in spinning disc, mirror in-the-lid, early color sets, 9AP4 picture tubes. Arnold Chase, 9 Rushleigh Road, West Hartford, Conn. 06117. (203) 521-5280.

TRANSMITTER, Hallicrafter BC-610-I, 2-18 MHz, 400W, good condition $225 or best offer. Glenn, 3748 Yosemite, San Diego, CA 92109. (619) 272-7538.

MOBILE IGNITION SHIELDING. Estes Engineering, 930 Marine Dr., Port Angeles, WA 98362.

DRAKE L4B with P.S. Pickup only $475.00. Wanted: Ten-Tec 2591. Hal, W6LXZ. (714) 679-7837.

CABLE CONVERTERS, decoders. Catalog $1 refundable. APS, POB 263 HR, Newport, RI 02840.

FOR SALE: HW-101 with CW filter, power supply, speaker, Shure 444 microphone. Mint condition, all for $350.00. Paul D. Dilliard, WA8IBT, 1313 East Wheeling St., Lancaster, Ohio 43130. (614) 653-5268.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners **RTTY** Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

RACAL RA.17 Military/Commercial grade communications receiver. The original up conversion design with continuous coverage from .5 - 30 MHz. Selectivity in 6 steps from 100 Hz to 13 kHz. North American model of British performance equivalent of Collins R390 without mechanical bandswitching. Complete set of replacement tubes. $450. John Harding, 1029 Sunset Drive, Blue Bell, PA 19422.

SATELLITE TELEVISION SYSTEM. Order the 2-year warranted Dexcel/Gould 1100-01 receiver/L.N.C. with precision 9′ spun aluminum antenna and ball-bearing polar mount at our call-for-current lowest price (approximately $1695.00) and we'll complete the system with a Chaparral Polarotor I/control, hook-up assembly, 125′ line and installation instructions and assistance. Videosat, Inc., (602) 445-0383, Box 449, Prescott, Arizona 86302.

"KT5B" MULTI-BAND Antenna 160-10m $59.95, SO239/PL259 weather boot kit $5.95, 2m DDRR mobile antenna $39.95, 2KW roller inductor 28mh $47.50 plus more! Kilo-Tec, P.O. Box 1001, Oakview, CA 93022. Tel: 805-646-9645.

HAM HOLIDAY Sri Lanka. Write to Spangles Travels, 84 Templers Road, Mount Lavinia, Sri Lanka. Enclose 5 IRCs.

COSMOPHONE 1000: Who can send me a copy of schematic diagram/manual of this old transceiver (1959)? Please write to DL9WW, Kurt Bergmann, Koenigsbergerstr. 62, 5100 Aachen, Germany.

HALLICRAFTERS $$$ Serious collector needs Hallicrafter and other Ham equipment manufactured before 1940 for restoration and eventual museum exhibit. Need Hallicrafters, National, Hammarlund, Patterson, RCA, RME, Grebe, etc. Condition not important. Also need QST mags Vols. I & II and old tubes. All letters answered. Write Dave Medley, WA5YXA, 6621 Duffield Drive, Dallas, Texas 75248.

GALAXY 5MK2 xcvr, p.s., manual, four extra finals, Turner plus 2 mic, Hallicrafters T.O. Keyer, Vibroplex paddle $250. WB7NRE (401) 822-1943.

WANTED: Cash paid for used Speed Radar equipment. Write or call: Brian R. Esterman, P.O. Box 8141, Northfield, Illinois 60093. (312) 251-8901.

WANTED: Pre-1950 bugs for my collection. Vibroplex, Martin, Albright, Melehan, Boulter, etc. K5RW, 1158 Midway, Richardson, TX 75081.

WANTED: Old RCA, Western Electric tubes. (713) 728-4343. Maury Corb, 11122 Atwell, Houston, Texas 77096.

PARABOLIC ANTENNA, spun aluminum, 6 ft. with mount $325.00. 408-730-2500. Norman, 2225 Sharon Rd., #224, Menlo Park, CA 94025. (415) 854-0266.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

WANTED: Boonton/HP type 250B RX meter also service manual for ex USN receiver type R1051B/URR. Write: Gill, 72 Elgin Street, Gordon, 2072, Australia.

COLLINS KW-1 parts wanted. Chuck, Box 766, Dahlgren, Virginia 22448.

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WE GOT OUR CHRISTMAS PRESENT — Super reviews on "MUFPLOT," a band opening predictor for Commodore computers. Get yours at a good price! Free details from Base(2) Systems, 2534 Nebraska, Saginaw, Michigan 48601. (517) 777-5613.

INDEPENDENT REP. ORGANIZATION wishes to sell out or merge. Good knowledge of two-way Radio systems and test equipment necessary. Sales ability a must. Top notch lines. Territory all of New England. Respond to: Box O, Greenville, NH 03048.

Coming Events

ACTIVITIES

"Places to go..."

ILLINOIS: RA-COM '83 sponsored by the Mt. Prospect Amateur Radio Club and Tri-County Emergency, November 13, Prospect High School, 801 W. Kensington, Mt. Prospect. Doors open 8 AM. Large indoor flea market, exhibits, seminars and more. Talk in on 146.52. For information, flea market or booth reservations: SASE to RA-COM, P.O. Box 452, Mt. Prospect, IL 60056.

MASSACHUSETTS: The Honeywell 1200 Radio Club, sponsor of 147.72/12 repeater and the Waltham Amateur Radio Association, sponsor of 146.04/64 repeater, will hold their annual Amateur Radio and electronics auction, Saturday, November 19, Honeywell Plant, 300 Concord Road, Billerica. Exit 27 off Route 3. Doors open 10 AM. Free admission and parking. Snack bar and bargain parts store. Talk in on both repeaters. For information: Doug Purdy, N1BUB, 3 Visco Road, Burlington, MA 01803.

MICHIGAN: The Oak Park High School Electronics Club presents the 14th annual Swap 'N Shop, Sunday, November 27, Oak Park High School, Oak Park. 8 AM to 4 PM. Admission $2.00. 8 ft. tables $6.00. Refreshments available. For information/reservations: SASE to Herman Gardner, Oak Park High School, 13701 Oak Park Blvd., Oak Park, MI 48237. (313) 968-2675.

MINNESOTA: The annual Handi-Ham Winter Hamfest, Saturday, December 3, Eagles Club, Faribault. Registration 9 AM. There will be a Handi-Ham equipment auction, dinner at noon followed by a program. Talk in on 19/79. For information: Don Franz, WØFIT, 1114 Frank Avenue, Albert Lea, MN 58007.

NEW YORK: Radio Central ARC presents the 5th annual "Ham-Central" All inside flea market and Hamfest, Sunday, November 27, Temple Isaiah's main social hall, 1404 Stony Brook Road, Stony Brook, Long Island. Doors open 7:30 AM for sellers/dealers; 9 AM general admission. Tickets $3.00 (OM or XYL and kids under 12 free). $7.00 for 8 ft. table space includes one free admission. Free parking. Nearby shopping. For information or reservations: Scotty Policastro, KA2EQW (516) 589-2557, 80 7th Street, Bohemia, NY 11716 or Bob Yarmus, K2RGZ (516) 981-2709, 3 Haven Ct., Lake Grove, NY 11755.

NORTH CAROLINA: The Guilford Amateur Radio Club's annual Hamfest/Computerfest, November 26 and 27, National Guard Armory, Greensboro. 9 AM each day. Admission $3.50 advance; $5.00 at gate. Tailgating allowed with price of admission. Food and free parking. Talk in on 144.65/145.25 and 146.52 simplex. An equipment check-out booth with test equipment and a technician available free for those wishing to check equipment prior to purchase. For information or advance tickets SASE to GARC, P.O. Box 7007, Greensboro, NC 27407. Please make checks payable to GARC.

OHIO: The Great Lakes/Ohio Valley Satellite Technical Show and Consumer Fair will be held November 19-20 at the University Hilton, Columbus, Ohio. Brought to you by Satellite Reception Systems, Athens, Ohio. Consumer Ticket Price $3.50. 1-800-592-1956 National, 1-800-592-1957 in Ohio.

OHIO: The Massillon Amateur Radio Club, W8NP, will present "Auctionfest '83" Sunday, November 13, 8 AM to 5 PM, Massillon K of C Hall, 988 Cherry Road N.W., Massillon. Flea market setup 7 AM; auction at 11 AM. Advance tickets $2.50, $3.00 at door. Tables $5.00 per 8 ft. space. Talk in on 147.78/.18. For information and reservations SASE to MARC, 920 Tremont Avenue S.W., Massillon, Ohio 44646.

PENNSYLVANIA: The Foothills ARC's 15th annual Hamfest, Saturday, November 5, St. Bruno's Church, South Greensburg. Tickets $2.00 or 3/$5.00. Indoor flea market tables $5.00. Mobile check-in on 146.07/67. For information, tickets or tables: WA3HOL or write FARC, P.O. Box 238, Greensburg, PA 15601.

OPERATING EVENTS

"Things to do..."

NOVEMBER 3 TO 6: The NBS-BRASS of Gaithersburg, Maryland, will operate K3AA to observe the dedication of the first active Amateur Radio Club station at the National Bureau of Standards. Multi-op activities on CW, Phone and RTTY near low end of 80 to 10 meter Novice and General class bands. Certificate available for SASE to BRASS, c/o National Bureau of Standards, Mailroom, Washington, DC 20234.

NOVEMBER 5 AND 6: Radio Central ARC, Rocky Point, New York, will operate WA2UEC from the former RCA HF Radio station called "Radio Central" to commemorate the 62nd year of the now silent station. 2-160 meters up 10 kHz from edge of General band and on 2 meters on 146.52 and 144.550/145.150 repeater. Novice band operation 7.110 kHz. For a special QSL card showing a photo of the former station send your QSL with large SASE to Radio Central ARC, P.O. Box 680, Miller Place, NY 11764 or QSL to Callbook address.

NOVEMBER 11, 12 AND 13: The Armored Force Amateur Radio Nationwide Emergency Team (A FAR NET) will help commemorate Veteran's Day by operating a special event station, 1200 UTC and 2400 UTC on all three days. 40 meters: 7280 to 7290 kHz. 20 meters: 14,320 to 14,330 kHz. 15 meters: 21,370 to 21,380 kHz. Those making contact with member stations can obtain a commemorative certificate by sending $1.00 to Harry B. Thomsen, W2PJH, 348 Jefferson Avenue, Apt. 15, Canadaigua, NY 14424. Indicate call letters of station contacted, the station's A FAR number, date, time and bands. Include call, name and address.

NOVEMBER 24: Thanksgiving Day. A special events station sponsored by the Whitman ARC and Plimoth Plantation will operate from the Plimoth Plantation's 1627 Pilgram Village museum. Call WA1NPO, 1300 GMT to 2000+ GMT. This event will be supported by members of the Plymouth (Devon, England) Radio Club operating G3PRC from a site overlooking Plymouth Sound from which the original "Mayflower" set sail in 1620. To receive a certificate, send proof of contact and $1.00 or 3 IRC's to: Whitman ARC, P.O. Box 48, Whitman, MA 02382. For additional information: KA1CZS (617) 826-4772; WB1CNM (817) 586-7524; Rosemary Carroll, Plimoth Plantation, P.O. Box 1620, Plymouth, MA 02360 (617) 746-1622; or Peter Jackson, G3ADV, 32 Brown Avenue, Parkfield, Nantwich, Chesshire, UK, Phone 0270-626149.

NOVEMBER 26 TO JANUARY 6: The Niagara Falls Radio Club will operate special event station W2QYV during the Festival of Lights from Niagara Falls, New York, 1500 UTC to 0300 UTC in the General portion of 20, 40 and 80. For a color photograph award send QSL and $2.00 donation along with 8½ × 11 SASE (55¢ postage) to Angelo Zino, WA2UJR, Awards Manager, 16 Council St., Niagara Falls, NY 14304.

DECEMBER 3: The Connecticut DX Association will operate KO1R, 1300 to 2000 Z, from the home of Mark Twain, Mark Twain Memorial, Hartford, CT. Frequencies for Phone and CW will be lower portion of General and upper portion of Advanced bands. For a full color QSL send your QSL and SASE to: Conn. DX Assn., P.O. Box 181, Columbia, CT 06237.

DECEMBER 3 TO 5: The Telephone Pioneer Radio Amateurs, John D. Burlie Chapter No. 89, 19th annual QSO Party. T.P.A.R. Operators in U.S. and Canada are invited to participate. Starting 1900 UTC, Saturday, December 3 to 0500 UTC, Monday, December 5. Suggested phone frequencies: 3.895-3.935; 7.255-7.295; 14.265-14.305; 21.355-21.395; 28.685-28.725; 50.10-554.00; 144.100-148.00. Contacts via repeater or simplex are valid. Suggested CW Frequencies: 3.555-3.595; 7.055-7.095; 14.055-14.095; 21.055-21.095; 28.055-28.095; 50.00-54.00 (Novice/Tech): 3.725, 7.125, 21.125, 28.125. RTTY use customary freq. Scoring: contact points times chapters contacted. Only one multiplier taken for each chapter worked. Maximum multiplier is 98 (TPA chapters 1-98) plus no more than 10 ITPA chapters. Exchange: Contact and chapter numbers. Return log sheets via your ARC Coordinator showing date, time, station worked, chapter name and number, contact number and claimed score. Post-marked no later than January 15, 1984. Send to: Ted Phelps, W8TP, John D. Burlie Chapter No. 89, Telephone Pioneers of America, c/o Western Electric, Dept. 45430, 6200 East Broad Street, Columbus, Ohio 43213.

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contents

presstop de W9JUV

AN UNPRECEDENTED FCC-AMATEUR ANTI-INTERFERENCE PROJECT was kicked off in Chicago October 19 by the FCC Chicago Field Office. Introduced to about 70 area repeater representatives by the plan's originator, new Chicago Engineer-in-Charge Joe Monie, WBØPAW, the "unconventional approach" program will be conducted strictly on a trial basis to see if such FCC-Amateur cooperation is workable. Though strictly local in scope, the project is being conducted with the knowledge and agreement of appropriate FCC Washington bureaus.

Initial Thrust Of The Experiment Is Against Intentional Interference to area 2-meter repeaters. Several Chicago-area machines have long been plagued by this problem, and the project was initially conceived in response to their specific difficulties. All liaison between FCC and participating Amateurs will be conducted under strict guidelines, with very specific criteria as to what types of information will be acceptable for FCC use in hunting down and eliminating interfering stations.

Reaction To The Proposed Program Was Quite Positive, with some urging that its scope should eventually be expanded to cover other kinds of interference and other bands. If it is successful Monie feels that it could also be adopted by other FCC Field Offices as well. Though at this time the project is not a part of the long-planned Amateur-FCC enforcement program made possible by the Goldwater Communications Act rewrite bill, it could well become a significant aspect of it if the project does prove workable.

A Fringe Benefit Of The FCC-Sponsored Meeting was the reactivation of the long-dormant Illinois Repeater Council. All Illinois repeaters should contact WD9GMZ to register their interest in participating in future IRC activities.

GRENADA'S POLITICAL UPHEAVAL WAS GRAPHICALLY BROUGHT HOME to listening Amateurs by KA2ORK/J3A, a student at St. George's University School of Medicine, who came on shortly after military operations began early October 25. Though Mark was able to confirm no U.S. medical students on the island appeared to be casualties, he reported both artillery and small arms fire plus frequent overflights of helicopter gunships and other military aircraft. Sometimes operating while prone on the floor of his shack to avoid possible stray bullets, he and J37AH provided the only real-time reports of the invasion action.

Interested Amateurs Were Joined By Other Amateur Stations with inputs to various government agencies and news media on 14250, 14303, 21300, 21309, and 21375. Many found themselves featured on TV and radio, as normal news sources were unavailable.

STS-9'S OCTOBER 28 LAUNCH WAS SCRUBBED due to severe erosion of launch vehicle rocket nozzles, with consequent delay in W5LFL's operation from space. Though an alternate date of November 28 seems probable at press time, it's also possible that the nozzle problem or other considerations might further delay the STS-9 launch to February 22.

Unfortunately, Limitations On W5LFL's Operating May Be Much Greater than originally expected. Unless the mission's operating schedule is changed, only nine orbits will have the proper combination of spaceship orientation and W5LFL's free time to put his signal into the U.S. In addition, most of these orbits favor the east coast. Unless there's a further change, he'll be active on orbits 49, 56 (descending), 64, 80, 96, 112, 113, and 129. It's also possible that he may be able to operate during some additional orbits, or that the radiation pattern from the spaceship may permit contacts even when his window antenna is oriented away from the earth. Check the phone numbers in November Presstop for updates, but note that Westlink's number for STS-9 news is now 213-465-1500.

THE 10-YEAR AMATEUR LICENSE BECAME A REALITY OCTOBER 6, when the Commissioners acted favorably on PR Docket 83-337. At the same time they also extended the grace period for station license renewal from one year to two (operator licenses already are five years).

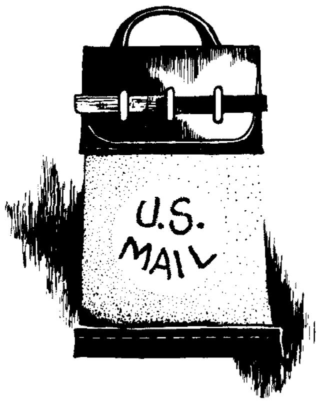
Current Licenses Will Not Automatically Be Extended. Instead 10-year licenses will be phased in with renewals. Just when the new 10-year licenses will actually begin to be issued depends on when new computer programs can be implemented.

TECHNICIANS MAY RUN FOR ARRL OFFICE, League directors decided at their fall meeting in Houston. The previous requirement was for candidates to be General Class or higher.

FCC's Volunteer Exam Program Was Endorsed By The Directors in a unanimous resolution, which also specified, however, that League participation in the program would commence only "...upon governmental authorization for recoupment of VEC expenses..."

220 MHz CONTINUES TO ATTRACT POTENTIAL USERS, with Waterway Communications System, Inc., pushing its FCC Petition for Rule Making (RM-4560) for a dedicated inland waterways band from 216 to 220 MHz. In addition, the FCC's Private Radio Bureau has also issued a "Final Report" on a study by its planning staff on future land-mobile needs that urges "Reallocation of 2 MHz from the 216-225 MHz band to the private land-mobile radio services for narrowband systems using 5 kHz spacing." RM-4560 is being strongly opposed by the Association of Maximum Service Telecasters because of possible channel 13 interference.

The 420-450 MHz Band May Also Be Hearing More Non-Amateur QRM soon. Westlink reports that Hughes Aircraft has a new contract for a military "Position Location and Reporting System" in that band for use by the Army and Marine Corps.



comments

power supply

Dear HR:

WB2UAQ's comment (July, 1983) on the March, 1983, article, "Dual Voltage Power Supply," suggested replacing the LM317 with a 723-type regulator. I did just that (see **fig. 1**).

Paul B. Johnson, VE7DHM
Sooke, British Columbia

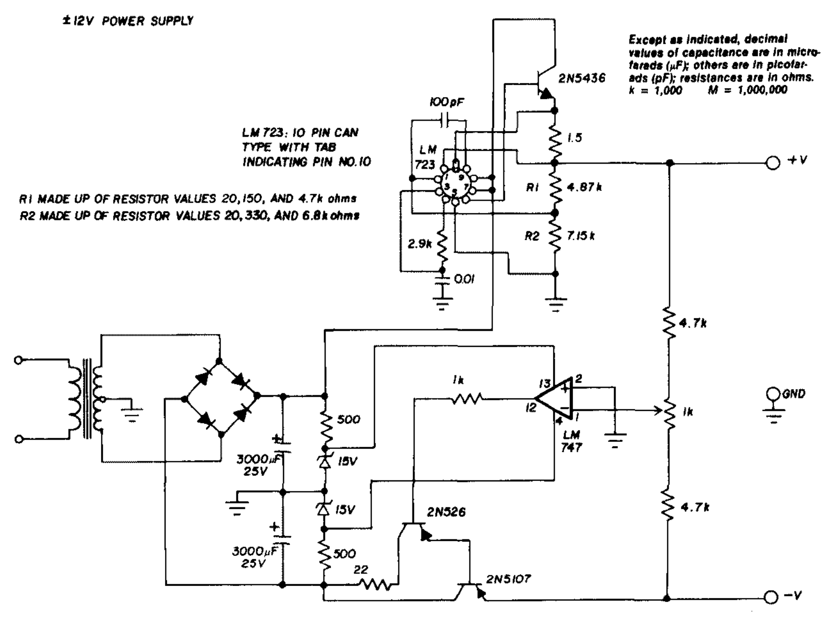


fig. 1. Dual voltage power supply uses an LM723 regulator.

loaded antennas

Dear HR:

Bill Orr's column on loaded antennas (*ham radio*, April, 1983) impressed me because although I generally operate on 20 meters, I'd like to use the 80-meter band and would like to have an antenna that doesn't require a ground.

Using the information on pages 54 and 55 of that issue, I wound the 40 μH loading coils but used No. 14 instead of No. 16 wire and checked the results on a Heath inductance bridge. Although the inductance was just about at the bottom of the scale (the bridge operates at 100 Hz, I believe) the results were accurate to two places.

I could not measure the center inductance but wound 11 turns of No. 14 wire on a 3/8-inch danvas impregnated form I had previously used for RF work.

It's OK to use the inductances *if* you have the equipment to measure them. Perhaps it would be better to indicate the coils not in inductance value, but rather in frequency, using a 10 pF capacitor to resonate with it. (Mine resonated at 14.25 MHz using a Millen meter for measurement.)

My friend WB6AFJ has an LC loaded 80-meter antenna. He's had plenty of trouble with the capacitors burning up. The absence of the capacitor, and the high voltage developed on these coils makes them difficult to manufacture for powers up to the legal limit which he uses now and then, as I do. I tend to shy away from coils as a result, but these — without the loading capacitors — look good.

One thing, however. Using the old compression insulators in which the wires overlap, how does one calculate the full dimensions of the lines? I used small ceramic insulators, and made the wires to full dimension plus about three feet; these I folded back and secured. I like the adjustability of this method. The doubled-back lengths may add a little capacitance on the ends of the wires, and may assist in loading. The antenna has not been dipped as yet; I have to get the noise bridge out and adjust it to specs.

I did not attempt to dip the 1.86 μH coil, but I ought to. I may have to use 52 ohm coaxial; right now I have RG-59 which is 75 ohms instead of 52 ohms.

Someone might want to run a program showing the resonant frequency of the coils mentioned in **fig. 4** (page 54) for benefit of those using the figures and not having other equipment.

Hank B. Plant, W6DKZ
San Jose, California

clean it up

Dear HR:

In his Ham Note, "Low Duty Cycle Transmitter Tune-up," (August, 1983), K4KI recommends using an automatic keyer in the dot mode during transmitter tune-up in order to reduce the duty cycle by approximately 50 percent and thereby save wear and tear on the tubes. But although an editor's note recommending the use of a dummy load at all times was included, we all know not every ham has a dummy load. Those who do have dummy loads don't always use them, and there are times when loading must be done into a radiating antenna, even if only briefly.

Therefore, K4KI's alternatives of either feeding the keyer's audio sidetone through the microphone, or feeding in audio generated by a relay connected in an RC time-constant circuit to make it buzz were both most unfortunate suggestions. Hopefully anyone currently using either of these techniques while loading into anything other than a dummy load will discontinue the practice.

While it is true that a pure sine wave carefully fed into the microphone input of an SSB transmitter can produce, for all practical purposes, a sine wave RF carrier in the output, this should never be attempted casually. The signal should be monitored on an oscilloscope, as the slightest distortion of the input sine wave will result in trash signals in the RF output which may vary from mild to an RF carrier output composed of numerous signals. Add in some flat-topping in the finals, and this garbage will extend perhaps several hundred kHz above and below the intended frequency of operation.

Few, if any, audio sidetones are pure sine waves. This is true whether it is electronic keyer sidetone, or the CW sidetone audio now included in all modern transceiver designs. At least one commercial electronic keyer I've seen uses a diode in series with the speaker, clipping one half of the audio waveform and generating a truly unique sidetone signal! Sidetone signals also tend to have an abundance of clicks or chirps, often both. Many of the CB to 10-meter conversions have no provision for operating CW if they are ex-CB SSB transceivers. It is common to feed keyed sidetone audio into them through either the mike or mike input circuit, and the result is spectacular. A number of these are currently loose on 10 meter CW and their signals are characterized by what can only be described as sounding like keyed steam calliopes — numerous carriers, usually accompanied by a bad case of chirps and/or clicks. A fairly clean one will occupy 5 or 10 kHz.

Audio sidetone-generated signals fed into the mike or mike input circuit will invariably generate unsanitary RF output from an SSB transmitter. Sending it into a dummy load is one thing, but radiating this garbage is quite another. Where on-the-air transmitter tuning is unavoidable, only the CW mode utilizing the transmitter's internal CW keying circuit and a keyer set for fast dots is appropriate. Hopefully the internal keying circuitry will provide proper shaping to avoid generating key clicks (approximately 5 ms rise and decay times), and if the dot/space ratio is correct, this will reduce the duty cycle to something less than 50 percent.

Robert G. Wheaton, W5XW
San Antonio, Texas

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short circuit

phased verticals

One line of K2BT's article, "Phased Vertical Arrays: part 4" (October, 1983) was inadvertently omitted. The second-to-last sentence on page 45 should read as follows:

"The calculation procedures are structured and identical for any circuit (except for the differing equations for matrix values), making the method ideal for programmable calculators or small computers."

Note: the front cover illustrates state-of-the-art techniques in the reception and processing of ionospheric sounding returns. The color-enhanced display was recorded by a *Digisonde* installed at Goose Bay, Labrador. It combines in one ionogram both vertical and oblique echo returns where: *yellow* signifies vertical echoes with ordinary polarization; *green*, vertical echoes with extraordinary polarization; *blue*, oblique echoes; *magenta*, automatically identified ordinary F trace; *red*, automatically identified ordinary E trace and extraordinary F trace. (Photo courtesy of Professor Bodo W. Reinisch, Technical Director, Center for Atmospheric Research, University of Lowell, Massachusetts.)

digital ionosondes

A second-generation probe of the ionosphere

HF radar is practically a household word to Radio Amateurs these days with almost everyone familiar with the Russian woodpecker and some with the quieter U.S. equivalent installation in Maine. These are known as Over-The-Horizon Backscatter Radars or OTH-Bs for short. Their purpose is one of long range aircraft detection. Another type of HF radar, known as ionospheric sounder, also operates using the echo principle. The basic differences between these two radars are range and type of object detected. While both radars utilize ionized layers, OTH-Bs use them as a means to an end, while sounders concentrate on the layers themselves. Information provided by the sounders enables communicators to design radio systems, choosing frequencies and times of operation more effectively. It is this latter type of probe that will be discussed.

Sweep-frequency ionospheric sounding equipment was first developed more than 50 years ago,[1] but until recently there was only one type of sounder or ionosonde. It operated in an analog mode, in the 1 to 30 MHz frequency range, by transmitting pulses vertically or obliquely to the D, E, and F ionospheric regions and receiving the weaker returning signals. The round trip time of the signal determined height, while the intensity of the echo could be related to the degree of ionization of the D region(s), where radio wave energy is absorbed while signals propagate through it to the higher reflecting layers. Though extremely useful data became available in the form of ionograms (see **fig. 1**) it still represented a two-dimensional display in a multi-dimensional world. What was needed was instrumentation that determined not only amplitude but also phase, direction of arrival, and polarization of the signal returns.

The next logical step in sounder development took advantage of the high speed capabilities inherent in digital data processing techniques. By combining a general computer with a specialized RF processor, a digital ionosonde was created. It now had the capability of real-time data analysis and display by processing the (complex vector) numerical description of signal returns as they varied with range and frequency.[2] Manipulation of this data provided additional signal parameters including phase, group delay, and doppler spectrum. When combined with an array of receiving antennas, the digital ionosonde was also able to determine signal direction of arrival, wave polarization, and other diffraction pattern information. The significance of this development lies in the fact that a single piece of equipment could now provide all this information, whereas in the past, several different equipment systems — along with elaborate measurements — were necessary. The insight provided enabled ionospheric researchers to re-examine basic concepts in light of this new information.

ionospheric regions and layers

December 12, 1901, marks the date when the first successful reception of transatlantic signals occurred. It also marks the start of inquiries into the mechanism that allowed this radio transmission around the curvature of the earth. A.E. Kennelly from the United States and O. Heaviside (remember the Heaviside layer?) in Great Britain independently conceived of the existence of an ionized region in the upper atmosphere capable of reflecting radio waves back to earth.[3,4,5]

The ionosphere, a term first used by Sir Robert Watson-Watt, was defined as "that part of the atmosphere in which free ions exist in sufficient quantities to affect the propagation of radio waves."[6]

By Rich Rosen, K2RR, Editor-in-Chief, *ham radio*

(However, when Sir Edward Appleton looked into the physics of radio wave propagation in a plasma, he found that the refractive index of the plasma, which controls the propagation of the wave through it by changing the phase velocity of the wave, depends mainly on the electron density of the plasma. The ions, because of their large mass — compared to that of an electron — have little impact on wave propagation. But since charge neutrality requires that the number of ions equal the number of electrons, it could be said that Sir Robert was partially right.) It is accepted as existing from approximately 31 miles (50 km) to as high as several earth radii. There are three commonly-known sections of the atmosphere called the D, E, and F regions occurring at heights of 31 to 56 miles (50 to 90 km), 56 to 87 miles (90 to 140 km) and above 65 miles (120 km), respectively. (The regions are not clearly defined and merge with one another.) These regions can also be divided into smaller layers of ion distributions, with the E region occasionally showing E1, E2, and Es layers while the F region divides into F1, F1½, and F2 layers.

The F1 layer has a maximum between 160 and 180 km, exists only in the presence of sunlight, and has a maximum density at local noon. The F2 layer peaks between 200 and 600 km, depending on factors such as time of day, season, phase of solar cycle, neutral winds, ion composition, etc. Due to the low densities at these altitudes, recombination (electron/s + ion = neutral) is very slow; the ionization exists for many hours after sunset. The F1½ layer occurs sometimes after eclipse events, but rarely under normal conditions.

The F2 layer is the most important layer for radio communications, since it generally has the largest electron densities and therefore reflects the highest frequencies; it is found at the greatest height, and therefore results in the largest possible 1-hop distance.

Some claims have been made for the existence of two other regions: C and G. The C region is thought to exist at the bottom edge of the D region, approximately 60 kilometers up, and is formed by cosmic rays and is therefore always present (since impinging cosmic rays are always present). The G region appears on ionograms as a little kink during a storm when the critical frequency of the F2 layer is greatly diminished. It's possibly not a distinct region but rather a phenomenon that occurs at special times only.

ionospheric terminology

To understand the role these layers play in communications it's necessary to define a few related terms. *For radio communications the most important* characteristics of the ionospheric layers are their *critical frequencies* and *virtual heights*. If one were to slowly vary the frequency of a generator that was transmitting pulses vertically (straight up) to the E region, a frequency would be reached where the signal was no longer being reflected down. The highest frequency that reflects back down from the E region is known as the E-region critical frequency or foE for short. Similar terms can be defined for the F1 and F2 layers: foF1 and foF2. Note that the subscript zero in the terms for critical frequencies represents zero distance (surface separation) between a pulse's origin and return location.

Virtual height is the equivalent height of the layer based on length or round trip time and apparent uniform velocity. In actuality, due to the interaction between the radio waves *and the electrons, a radio* wave is slowed down as it enters ionospheric layers, resulting in non-uniform velocity. Consequently, actual layer height differs from virtual height. Nevertheless, virtual height represents an accepted and useful convention as we shall soon see.

Up to now vertically launched signals have been considered. If the same transmitter drives an antenna that produces a *low* angle (elevation) pattern, additional useful information becomes available. Let us define MUF (2000) E and MUF (4000) F as the highest frequencies which the E or F layer will carry over a 2000, 4000 km path, respectively. These low angle terms when *divided by their respective layer critical* frequency equal what's known as the M-factor, i.e.

$$M(2000)E = \frac{MUF(2000)E}{foE} \text{ and}$$

$$M(4000)F = \frac{MUF(4000)F}{foF2}$$

Note that the F layer is not often high enough to give a 4000 km 1 hop. Knowledge of these sets of frequencies is one of the beginning steps in determining communication paths and the corresponding operating frequencies. This is where sounders come in.

analog sounders

An early form of an ionosonde was a pulsed radar device in which the frequency was repetitively varied from approximately 1 MHz to 30 MHz (some started even lower, at 0.1 MHz). The equipment was designed to measure the time it took a pulsed radio wave to travel up to the ionosphere and back as a function of frequency. The transmitter and receiver were synchronized either electronically or mechanically while the receiver output was displayed on a CRT. Markers were introduced that interrupted the sweep trace every one third of a millisecond and multiples thereof to provide equivalent height indicators of 50, 100, 150, . . . kilometers. (These are indicated

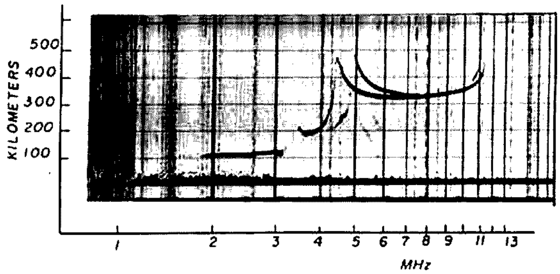

fig. 1. Vertical ionogram generated by an analog sounder on April 23, 1975, over Boulder, Colorado. Sporadic E is evident as well as ordinary and extraordinary wave F_1 and F_2 layers.

as thin horizontal lines in **fig. 1** and in this case are multiples of 100 kilometers or 0.66 milliseconds.) In addition, frequency markers were generated that corresponded to each integer MHz. (Note that the spacing between vertical frequency marker lines is non-uniform and in this case, logarithmic).

A common ionosonde, designed in the mid-1950's for the International Geophysical Year, that operates along these lines is the National Bureau of Standards (NBS) C-4 machine, still used today. It consists of a 10 kW peak pulse transmitter and a wideband receiver. The pulse transmitter uses a 31 to 55 MHz swept CW oscillator that mixes with a 30 MHz pulsed oscillator to provide a 1 to 25 MHz pulsed output. This signal is then amplified in a broadband power amplifier. The receiver uses the same CW (swept) oscillator in a balanced mixer to mix with the incoming echoes down to a fixed 30 MHz IF signal for further amplification and processing. The same antenna in an analog sounder is normally used both for transmitting and receiving and should show a relatively constant resistive impedance over the entire frequency range. Typical antennas utilized are vertically-oriented delta loops, terminated rhombics or log periodics.

A panoramic display of the returning echoes is ac-

ionograms explained

The dark, basically horizontal line on the left side of the ionogram represents the E region (specifically the sporadic E layer — a thin layer at about the same height as the normal E layer which varies, unpredictably). Signals from between 2 and 3 MHz (2nd and 3rd vertical lines) have apparently been reflected from a height starting at 110 and ending at 120 km on this film strip.

The second set of curves on the same ionogram illustrates several important features of ionospheric propagation. The first curve from start to cusp represents the F1 layer, while the next continuous segment from cusp to cusp represents the F2 layer. *The cusps or rapid increase in "height" with small change in frequency corresponds to the previous layer's critical frequency and indicates a maximum layer electron density. Furthermore, a cusp indicates that the electron density profile has a vertical tangent, i.e., 'that the specific layer has a maximum. The F1 layer often has no discrete maximum; it is merely a ledge on the F2 layer electron density profile — an inflection of the trace, not a cusp. It corresponds to points where the radio wave penetrates the layer. At the critical frequency, the rate of travel of the incident wave is slow, producing the large* virtual *heights.*

The last curve also shows the F1 and F2 layers (though it's hard to see the F1 trace in this case). The reason for the existence of two traces is related to the fact that the transmitted wave splits into two separate waves (ordinary and extraordinary) under the influence of the earth's magnetic field.[7] (This is known as magneto-ionic effects.) This trace separation is important to all communicators because these two waves propagate differently: traveling at different velocities and being absorbed in differing amounts. Depending upon a transmitting station's geographical location, the transmitting antenna's configuration and orientation (polarization and direction) as well as a number of other factors determines which wave is launched and how strongly it is reflected *if at all.*

Note that the following critical frequencies can be picked off the ionogram and are approximately:

foEs = 3.5 MHz	(Sporadic E critical frequency)
foF1 = 4.4 MHz	(Ordinary wave F1 critical frequency)
fxF1 = 5.0 MHz	(Extraordinary wave F1 critical frequency)
foF2 = 11.0 MHz	(Ordinary wave F2 critical frequency)
fxF2 = 11.4 MHz	(Extraordinary wave critical frequency)

complished by applying a voltage to the X plate of the CRT that is a function of the sweeping transmitter frequency. The time base voltage is applied to the Y plates which equates to the virtual height of the layers (round trip time between transmission and reception). The actual receiver output and height pips are applied as blanking pulses to the CRT. In examining the display (see **fig. 1**) two different curved lines (O and X) are seen that rise with increasing frequency to a cusp at the critical frequency of the F layer. This makes sense since higher frequencies are reflected at greater (layer) heights until penetration occurs (see "ionograms explained").

A second type of analog sounder used FMCW, which provided a linear frequency swept transmitted signal. Use of CW gave high average power eliminating the need for high peak powers. It also permitted the use of very narrow receivers (100-200 Hz bandwidth), and was therefore less susceptible to interference from fixed frequency communication signals. The analog output of this FMCW system is a delayed and distorted (by the ionosphere) replica of the transmitted FMCW waveform and not an ionogram. A mathematical operation called a Fourier Transform (or spectral analysis) on the signal determines the delay of each frequency component, which is a measure of the reflection height. This output provides the information which is plotted and recorded as an ionogram. In general the main advantage of using FMCW versus pulse was in its reduced level of interference to itself and other users of the spectrum. However, pulsed systems had an advantage over FMCW types when multichannel receivers or multiple antenna receiving arrays were used. It's easier to switch between multichannel receivers or sample multiple receiving antenna outputs on a pulsed system. By using the positive features of a pulsed system and adding some software controlled equipment for signal enhancement the basic digital ionosonde was created.

digital sounders

A simplified block diagram of one of the few digital ionosondes (**fig. 2**) in the world, the Space Environment Laboratory model,[2] is shown in **fig. 3**. The heart of the system incorporates a 16-bit processor. Data is displayed on a 6-inch (15 cm) XY CRT that

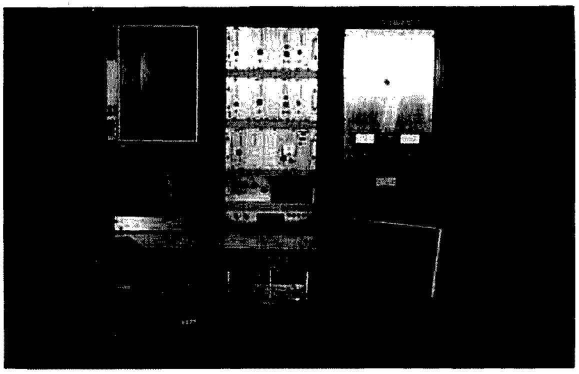

fig. 2. One of the few digital ionosondes in the world, this system developed by the Space Environment Laboratory provides new capabilities — within a single system — to determine ionospheric structure and magneto-ionic effects in general.

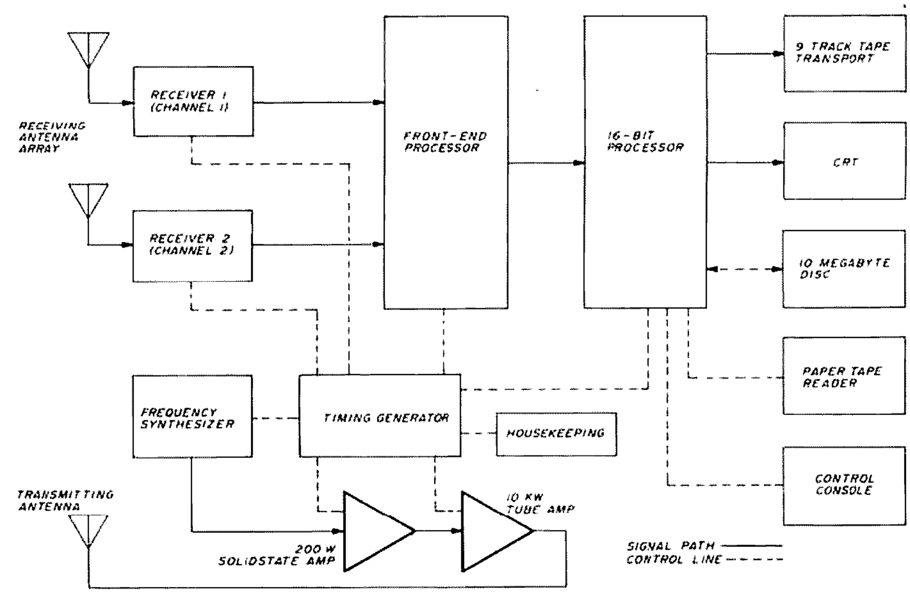

fig. 3. Basic block diagram of the Space Environment Laboratory digital ionosonde.

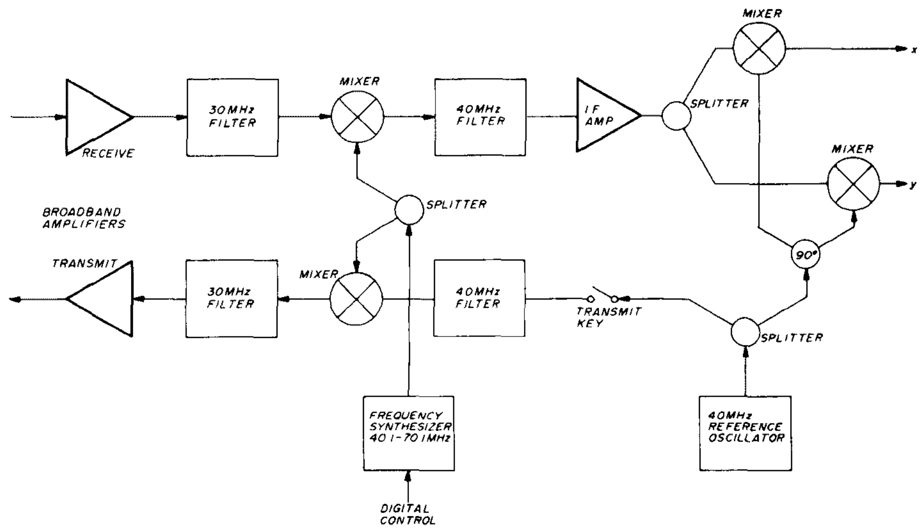

fig. 4. Simplified block diagram of the RF section of the SEL digital ionosonde.

has 1024 × 1024 addressable points. The display appears animated (is refreshed) by direct memory access to the computer's core. This makes rapid changes in the display possible and provides a time sequence presentation of the data. Program loading is provided by a 1600 BPI (byte per inch), 9 track, 3-ips tape transport while a 10 Megabyte disc memory is used for recording data. The operator talks to the system via a graphics display console and hard copy unit.

The RF section of the digital ionosonde has two functions: it generates the transmitted signals and coherently receives the echoes (see **fig. 4**). A symmetrical up/down conversion scheme is used with two oscillators. The first oscillator, a general-purpose synthesizer, generates a frequency between 40.1 and 70 MHz and up converts the receive band of 0.1 — 30 MHz to a 40 MHz IF. The second generator, a fixed-frequency 40 MHz crystal oscillator, mixes with the same synthesizer output to form the transmitted frequency.

In the receive mode the 40 MHz oscillator provides the reference signal while the frequency of operation is selected by the computer which controls the synthesizer.

An important question asked during the design phase was what is the best choice of output signal representation? Previous ionospheric sounding systems have, for the most part, used a logarithmic signal scale. The use of a logarithmic scale, however, is inconvenient for many kinds of digital signal processing. The justification for this choice of signal representation (as opposed to linear, for example) has been that it is otherwise difficult or impossible to achieve the received signal dynamic range any other way. Recent developments in solid-state technology have shown that stable, wideband linear amplifiers with extremely wide dynamic ranges, typically >140 dB in a 30 kHz noise bandwidth *are* achievable. Using these amplifier techniques, passive filters and a well-designed mixer and detector, a linear receiver can be built that is limited only by degree of (digital) quantization and DC stability.

Through careful design, the receiver section exhibits a high degree of linearity with low intermodulation products and achieves a 1 dB compression point of +15 dBm with a third order intercept of +26 dBm with 0 dB RF attenuation thrown in. In addition the receiver gain varies less than ±2 dB from 0.2 to 29 MHz and has a tangential sensitivity of ≤1 μV. An additional benefit of the strong signal handling performance of the receiver is that preselection filters are not required for operation under typical site conditions.

Two transmitted output levels can be used: 200 W or 10 kW. The low level transmitter drive output is amplified first by a solid-state class A amplifier to the 200 W level. This can be used directly or can be used to drive a pulsed class A wideband vacuum tube amplifier with a 10 kW output. Since the transmitter outputs are nominally 50 ohms, unbalanced wideband balun transformers must be used to drive typical sounding antennas.

operation

Since the key feature of this system is flexibility — i.e. it is software controlled and not hardware constrained — the digital ionosondes' modes of operation can be greatly modified almost at will. Initially, software has been written to operate the system as a

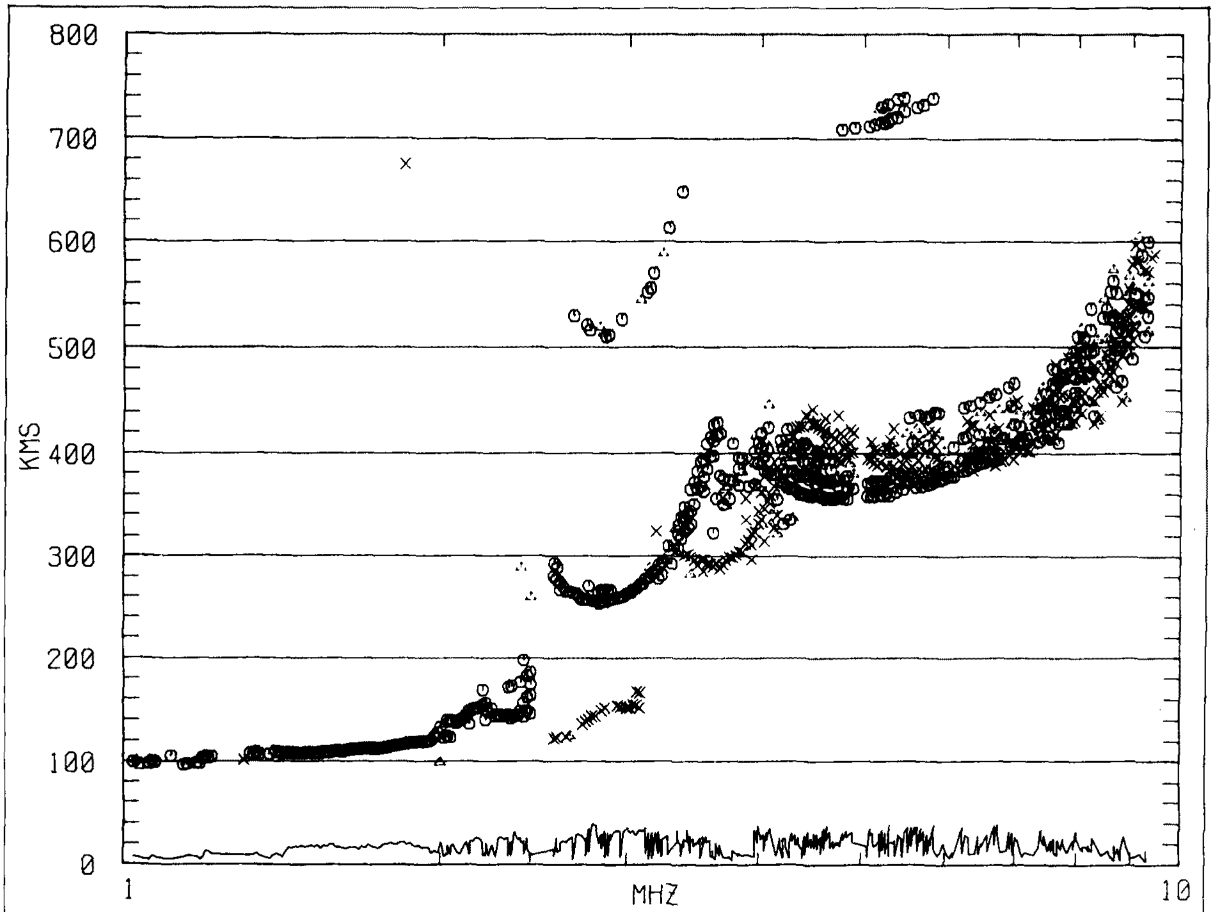

fig. 5. Digital ionosonde generated display depicting sporadic E, F1 and F2 layer ordinary wave returns as well as F1 and F2 layer extraordinary returns. The ordinary wave appears as red "O" symbols and the extraordinary wave as green "X" symbols on the display.

basic vertical pulse sounder, incorporating flexible sweep modes as well as fixed frequency operation (up to ten operator-selected frequencies.) Four pulses are transmitted per frequency, two at f and two at $f + \Delta f$ (Δf is usually chosen at 8 kHz). Four antennas are sampled, using two at a time with two receivers. The amplitude of the "returning signal" is examined (interference is removed) and transferred to the computer. After more processing, the data is recorded on tape and disc and displayed on the system CRT. Of particular interest to users of the standard frequency stations and Radio Amateurs is the frequency deletions of 2.5, 5, 10 MHz and the 160 and 80 meter bands by the system.

Fig. 5 shows a typical graphics display ionogram. The original, in color, shows the Es and F1 and F2 ordinary wave returns plotted as red letter O's while the F1 and F2 extraordinary wave returns are shown as green X's. The additional trace at the top is the second return trip echo (signal transmitted up to the F region, reflected back down, reflected by the earth, returned once again by the same F region. These are also seen on ionograms produced by the conventional analog sounders.)

digital ionosonde programs

In the Space Environment Laboratory Program in Boulder, planning calls for six of the SEL digital ionosonde systems to be built and coordinated as part of an international research program to primarily study the problems of magneto-ionospheric interaction in the Arctic and Antarctic ionospheres. Considerable software development needs to be done in order to take advantage of even a part of the possibilities of the system.

On the east coast, a series of digital ionosondes — the latest version known as Digisonde 256 — has been developed over the last decade by the University of Lowell[8] in cooperation with the Air Force Geophysics Laboratory at Hanscom Air Force Base in Massachusetts. The Digisonde is in use at military and civilian research and geophysical monitoring sta-

tions (Patrick Air Force Base, Florida; Ft. Monmouth, New Jersey; University of Lowell, Lowell, Massachusetts; Millstone Hill Incoherent Scatter Facility, Westford, Massachusetts; Air Force Geophysics Laboratory, Goose Bay Ionospheric Observatory, Goose Bay, Labrador; and at stations in the United Kingdom, Belgium, Germany and Italy). Additional sounders are being built for Australia, China, and France.

The Digisondes provide as routine station sounders digital ionograms[9] in the standard format (similar to **fig. 1**) measuring and displaying amplitudes, Doppler, angle of arrival and polarization of the ionospheric echoes in selectable combinations. These selections depend on station or experimental requirements. A special subsystem, the *A*utomatic *R*eal *T*ime *I*onogram *S*caler with *T*rue Height Analysis (ARTIST) scales the ionograms in real time and transmits via teletype ionospheric parameters and the complete trace defining the overhead ionosphere to users such as the Air Force Global Weather Center at Offutt Air Force Base in Nebraska, and the Air Force Over-the-Horizon Backscatter Radar at Columbia Falls, Maine. The system can also be operated in the so-called Drift-Mode, measuring ionospheric tilts and motions, and in a bistatic mode using two or more systems to provide propagation ionograms for communication link analysis.

acknowledgments

I thank Richard N. Grubb of the Space Environment Laboratory in Boulder, Colorado, for information on the NOAA SEL HF radar system. I also extend my appreciation to Jurgen Buchau of the Ionospheric Physics Division at the Air Force Physics Laboratory, and Bodo W. Reinisch of the University of Lowell, Lowell, Massachusetts, for information on the Digisonde 256 and their constructive comments on the material presented in this article.

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ham radio

state-of-the-art auto-dialer

It stores 21 telephone numbers, uses only 3 chips, and may save your life

Some time ago I designed a Touch-Tone® autodialer because I always had trouble dialing an autopatch call and driving at the same time. The dialer stored sixteen telephone numbers and was easily programmed from the keypad, but I had to use ten chips to make it work.[1] Now, because of a new chip on the market, I've found it possible to design a much better dialer that mounts directly on the dashboard. The new dialer has only three chips, and the parts were easy to get. To really simplify things, I even designed a printed circuit board. Anyone with a modest junkbox should be able to duplicate my autodialer for $50 or less.

features

The dialer has many features that make it attractive for Amateur use:

- It can store twenty-one telephone numbers, eleven of which can be twenty-one digits long.
- All programming and dialing is through a standard Touch-Tone® keypad; no other controls are needed.
- Any one of eight dialing speeds are available.
- Pauses can be programmed, with or without a carrier drop.
- An autopatch access code can be dialed at slow speed, followed by the telephone number at high speed.
- Manual dialing includes automatic PTT (push-to-talk) with a two-second hang-on time.
- PTT is keyed one second before any auto-dialed number starts.
- Audible feedback includes unique tones to distinguish between programming and dialing keystrokes. PTT is inactive during programming.
- Last manually dialed number can be redialed automatically.
- Battery backup retains memory and programming for five to ten years.

By Alan Lefkow, K2MWU, 17 Jacobs Road, Thiells, New York 10984

circuit description

A new CMOS Touch-Tone® auto-dialer chip called the MD-22 is the key to this design. It's available directly from the manufacturer[2] and comes in a 40-pin DIP package. The chip handles all housekeeping and dialing functions, including a standby mode for battery backup, the reading and writing of memory, tone generation, and a PTT output. It's used in conjunction with a standard 256x4 CMOS RAM for maximum number storage. To monitor tones, I added an audio amplifier chip and used an earphone as a loudspeaker.

For proper operation, the central part of the circuit follows the recommendations given in the data sheet for the MD-22. The schematic for the dialer is shown in **fig. 1.** Resistor network RN1 is an 8-bit R-2R ladder array. The network sums eight binary outputs from the MD-22 to produce a stepped approximation of a sine wave. (Most Touch-Tone® generator chips use the same approach, but the summing network is on the chip.) Capacitor C7 filters out the steps to produce a smooth waveform. Trim pot R4 acts as a volume control for amplifier U4, and R5 adjusts the tone output level.

Because of their low voltage drop, germanium diodes are used for CR4 through CR8. Silicon diodes are necessary for CR2 and CR3 because they carry higher current; their voltage drop is compensated for

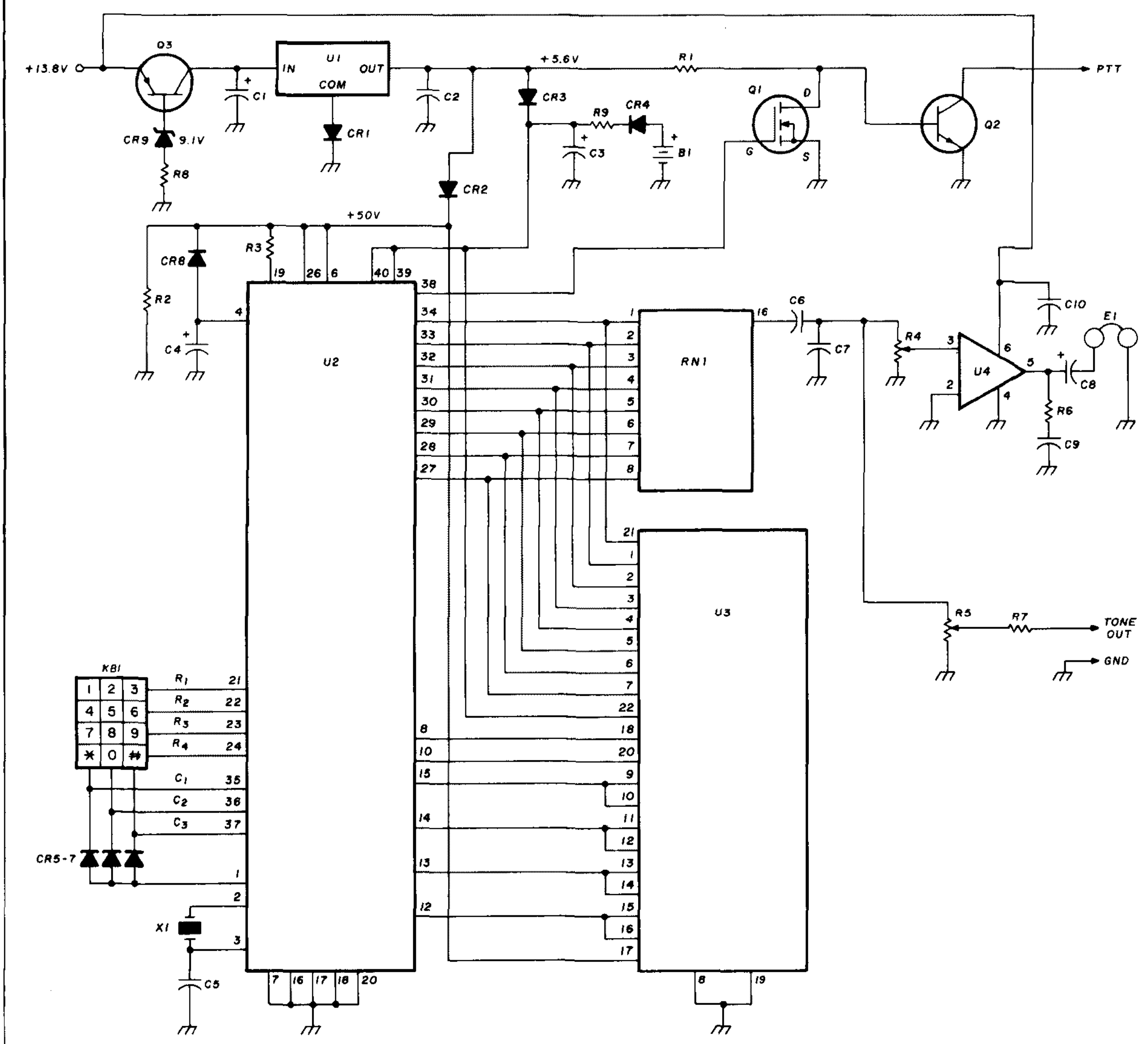
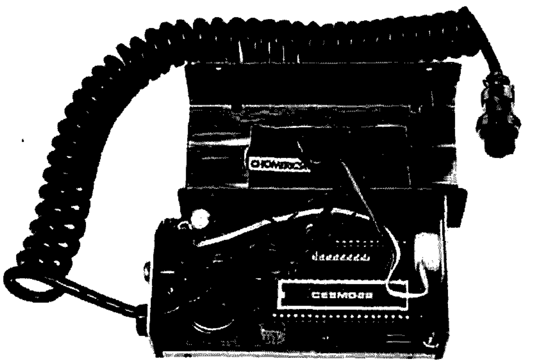


fig. 1. Three-chip autodialer schematic.



Inside view of the finished autodialer. The PC board shown here is an earlier design, slightly larger than the final version shown in *fig. 2*.

by using CR1 to raise the output voltage of regulator U1. This keeps the voltage for the MD-22 in its recommended range of 4.5 to 5.5 volts.

My first try at using the MD-22 pointed out some unexpected problems. The first one concerned the PTT output at pin 38. Most rigs require a switch closure to ground for PTT. Pin 38 goes low when PTT is on, but it must be buffered before connecting it to any rig. That I did with two transistors, using one as an inverter and the other as an open-collector output. Then *I discovered that pin 38 remained high* when the MD-22 was powered by the backup battery. That kept one transistor turned on and significantly increased the drain on B1.

To eliminate that drain, I followed a recommendation by Reiss[3] and changed the inverting transistor, Q1, to a VMOS FET. In this circuit the VMOS FET behaves much like a bipolar but draws negligible current from U2, which is just what I wanted. With this change, the drain on the battery during backup turned out to be less than 0.1 μA. At that rate, the battery should last five to ten years. When it does have to be replaced, the residual charge on capacitor C3 will provide ample time to replace the battery without any memory loss.

Another problem involved simply shutting off the unit. When power is removed, CR2, CR3, and CR4 switch in battery B1 to preserve memory. I discovered that if the supply voltage to the MD-22 doesn't drop fast enough when power is turned off, memory is erased. This can be a real problem if the autodialer is powered from a rig. The supply voltage doesn't drop abruptly when a rig is turned off because of large filter capacitors in the rig.

The manufacturer of the MD-22 explained the quirk. As the supply voltage falls to zero, there is one point when the backup battery is called upon to deliver current abnormally high for a small watch battery. This causes the backup voltage to the chip to collapse for a moment, and that erases memory.

The problem can be solved in one of several ways. The manufacturer recommended changing B1 to 4.5 volts and shorting R9 to provide higher current capacity. I tried that by adding a 1 ½ volt calculator battery in series with B1 and it worked, but I still preferred the compactness and simplicity of using a watch battery. That led me to add transistor Q3 with zener diode CR9 wired in series with its base. When the supply voltage drops below the zener voltage, Q3 turns off current to the dialer chip. The sharp knee in the diode's voltage curve switches off Q3 fast enough to prevent the loss of memory no matter how slowly the supply voltage falls. (On some samples of the MD-22 the Q3/CR9 combination may not work; if so, use the 4 ½ volt battery scheme.)

construction

You can make your own printed circuit board using the pattern in **fig. 2**, or you can order one from the source given in the parts list. Other construction techniques can also be used since parts layout is not critical.

If you use the board, the parts are installed following **fig. 3**. Use insulated wire for jumpers that might

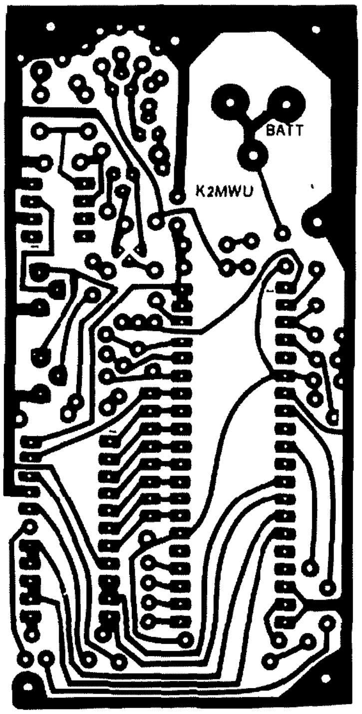


fig. 2. Printed circuit pattern for the autodialer. A drilled and plated PC board is available from the source indicated in the parts list.

parts list

item	*description*
B1	#2320 lithium watch battery (Radio Shack No. 23-163)
C1,C4	2.2 μF dipped tantalum
C2,C6,C10	0.1 μF
C3	100 μF dipped tantalum
C5	22 pF
C7	0.0039 μF mylar or polystyrene
C8	50 μF electrolytic or dipped tantalum
C9	0.05 μF
CR1-CR3	1N3600
CR4-CR8	1N270
CR9	1N757 9.1V Zener diode
E1	miniature earphone
K1	12-digit Touch-Tone® keypad, matrix switching
Q1	VN10KM VMOS FET (Radio Shack No. 276-2070 or similar)
Q2	2N2222A
Q3	2N3906
R1,R8	1K
R2,R9	10K
R3	5.1K
R4,R5	100K miniature trim pot, vertical mounting
R6	10 ohm
R7	75K or 2.2M (see text)
RN1	50K R/2R ladder network (Allen-Bradley No. 316L08503)
U1	78L05 5.0V low power regulator
U2	MD-22 Touch-Tone® autodialer
U3	5101 CMOS RAM
U4	LM386 audio amplifier
X1	6.0 MHz crystal, HC18/u holder (Jameco Electronics #CY6.00 or similar)
Case	LMB No. CR-425, 4-1/4 × 2-1/4 × 1-1/4 inches (L × W × H)

Except as noted, resistors are 1/4 watt, 5%, and capacitors are disc ceramic.

A drilled and plated printed circuit board is available for $7.50 per board, plus $1.50 for shipping up to 5 boards, from Dynaclad Industries, P.O. Box 296, Meadow Lands, Pennsylvania 15347. The MD-22 (U2) and 50K – R/2R network (RN1) are available for $15.00 and $2.50 respectively, including shipping, from CES Inc., P.O. Box 507, Winter Park, Florida 32790.

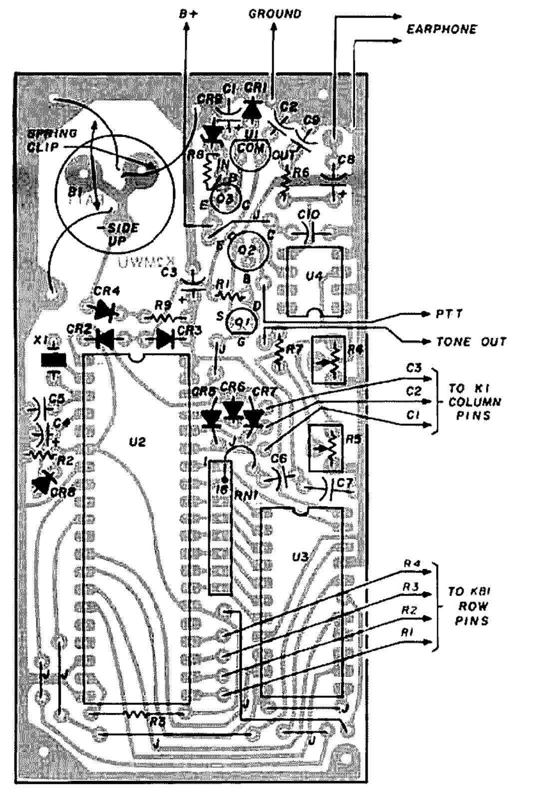


fig. 3. Parts placement for the PC board. All parts mount on the board except the keypad and earphone.

contact other components. And beware of static discharges when handling U2 and U3; they're both CMOS chips.

Most components mount as shown. However, the battery and resistor network require some additional work to install them on the board. The network comes in a 16-pin DIP package but only nine of its pins are used in this circuit. To simplify the PC pattern only one edge of the network is attached to the board. Pins 1 through 8 are soldered in place and bent such that the network is perpendicular to the surface of the board. Pins 9 through 15 are cut off, and pin 16 is connected to the board with a short jumper.

Fig. 4 shows how the battery is held in place using common nickel-plated safety pins fashioned into spring clips. The head and point of three pins are cut off and the pins soldered to the board. Some heat-shrink tubing should cover the portion of the pins closest to the board to keep the edge of the battery from shorting out to them. If desired, eliminating the middle pin at the corner of the board will make it easier to insert and remove the battery.

Another three holes on the board under the battery are used for soldering either eyelets, short pieces of bus wire, or small screws. They should project slightly above the surface of the board to make contact with the "+" side on the battery. The battery will be held against them by the spring clips.

Any standard 12-digit Touch-Tone® keypad that connects a row and column together whenever a key is depressed (called matrix switching) can be used for KB1. I used a discontinued Chomerics keypad purchased on the surplus market. The PC board was sized to fit neatly in the LMB Crown Royal case listed in the parts list. The earphone mounts in the small metal box too; just drill a hole and glue the earphone behind it.

interfacing to your rig

Other than the keypad and earphone, the only connections to the board are for power, ground, PTT and tone output. Power should come from a switched line in your rig that carries 12 - 15 VDC. The cable connecting the dialer to the rig should have a separate ground lead in addition to a shielded lead for the tone output line. Ground the shield at one end only, not at both.

Resistor R7 is used to attenuate the tones from the

dialer to match the requirements of your rig. Its value must be carefully selected. If too small, tones can get through to your rig due to crosstalk, even with R5 turned down to zero. This won't occur if R7 is large enough. When tones just begin to cause over deviation at the maximum setting of R5, R7 is at the right value. Use the # digit to check this because it generates the greatest deviation of any Touch-Tone® pair.

As examples, consider my two FM rigs. Touch-Tone® signals connect to the microphone input in both. In one, the 500 ohm microphone is always connected across the mike input, whether or not the PTT button is depressed. That means the dialer sees a load of 500 ohms. I found that 75K for R7 worked out just right. In my other rig the microphone impedance is 10K. There, the best value for R7 turned out to be 2.2 megohms. If your situation isn't covered by these two examples, experiment with R7, starting with a value that's 150 times the load impedance seen by the dialer.

programming and operation

Table 1 lists all the operations of the dialer and provides room for writing in programmed telephone numbers. Note that a # precedes every programming operation, while a * precedes an auto-dialing operation.

Let's cover manual dialing first. Pressing any key brings up the PTT line and outputs the Touch-Tones®, unless the first digit is a * or #. In that case, the * or # must be pressed *twice* to tell the dialer that no address or programming follows. The first of this double keystroke produces a special tone that's not transmitted, while the second one produces the normal * or # tones. After the last digit, the PTT line stays on for two seconds, at which time a warning tone sounds. (That tone won't be transmitted, either.) Any sequence of digits dialed before the warning tone can be redialed automatically by pressing * #.

Programming a telephone number is also simple. Key in the #, followed by the address, followed by the telephone number. Twenty-one addresses are available, consisting of the digits 1 through 9, and the combinations 01 through 0#, 00 excluded. You have two seconds to go from one digit to another before the warning tone sounds, indicating the end of programming for that number. (The end of the number can also be denoted by keying in # *, as shown on the first line in the table.) During programming, address digits produce special tones to distinguish them from the telephone number digits being stored. If the # is one of the digits to be stored, it must be keyed twice, just as before.

Any one of eight dialing speeds can be programmed from the keypad. Generally, telephone lines will accept the dialer's top speed, but many repeater autopatches need to be accessed at a slower rate. Rather than slow everything down, the dialer can be programmed to dial an access code slowly, and the telephone number fast. That's accomplished with the pause feature.

table 1. Chart of dialer programming instructions and list of addresses (telephone *numbers*) aids *mobile* operation.

	TO DIAL	TO PROGRAM	DIGITS																				
			1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21
	*#	LAST NO.	X	X	X	X	X	X	X	X	X	X	X										
	*1	#1												#* (OPTIONAL END)									
	*2	#2																					
	*3	#3																					
	*4	#4																					
	*5	#5																					
	*6	#6																					
	*7	#7																					
	*8	#8																					
	*9	#9																					
	*01	#01																					
	*02	#02																					
	*03	#03																					
	*04	#04																					
	*05	#05																					
	*06	#06																					
	*07	#07																					
	*08	#08																					
	#09	#09																					
	0	#0*																					
	*0#	#0#																					

MANUAL DIALING - IF FIRST DIGIT * OR #, PRESS TWICE

PROGRAMMED PAUSES FOR ACCESS CODES (TAKES ONE DIGIT STOPAGE)

#8	3 s.	PTT DROPPED	--
#9	3 s.	PTT DROPPED	UP-SPEED
#0	2 s.	PTT ON	UP-SPEED

PROGRAMMED SPEED:

SPEED	TONE ON	TONE OFF	RATE
#*1	80ms.	40ms.	8.3/sec.
#*2	80ms.	80ms.	6.3/sec.
#*3	160ms.	80ms.	4.2/sec.
#*4	160ms.	160ms.	3.1/sec.
#*5	300ms.	140ms.	2.3/sec.
#*6	300ms.	300ms.	1.7/sec.
#*7	600ms.	300ms.	1.1/sec.
#*8	600ms.	600ms.	0.8/sec.

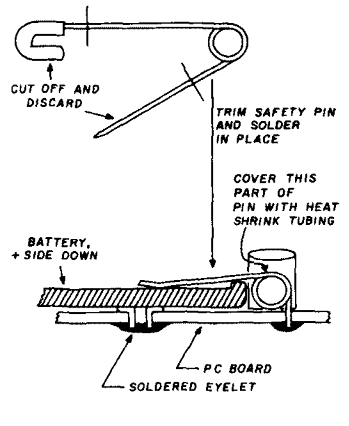


fig. 4. The watch battery is held on the board by spring clips fashioned out of ordinary safety pins.

As shown in the table, there are three types of pauses that can be progammed anywhere in a stored number. If the up-speed option is selected, the digits before the pause will be dialed at whatever speed the dialer is programmed for, but digits after the pause will go at top speed. The idea is to set the dialing speed slow enough to access the autopatch, and use an up-speed pause to send the rest of the number quickly.

When any number is autodialed, the PTT line is keyed one second before the number starts dialing. This also holds true when dialing resumes after a pause. This compensates for any delays introduced by mechanical switching in your rig or repeater.

Note that of the twenty-one addresses, eleven can store twenty-one digits and the rest, fewer. The table indicates how many may be stored for each. As a reminder, the dialer emits a unique tone when you enter the last digit an address can store. Since the dialer holds so many numbers, I suggest you make a photocopy of table 1, fill it in, and keep it near the dialer as a record of what you've programmed.

conclusion

The only real drawback to the MD-22 is the absence of the fourth Touch-Tone™ column. For most Amateurs that's not likely to be a problem. The real problem is driving with your eyes on the road and keying in ten digits to make a patch call. That's where the dialer serves its purpose well.

references

1. Alan Lefkow, K2MWU, "A Portable Touch-Tone® Auto-Dialer," *ham radio*, August, 1982, page 12.
2. CES Inc. See parts list.
3. R.A. Reiss, K1HOP, "Low-Power Keyer and Interface," *ham radio*, February, 1983, page 68.

ham radio

time and frequency standards: part 2

Multiple-frequency VLF techniques offer extreme accuracy

In part 1 (November) the atomic (non-quartz oscillator) frequency standards by which time and frequency measurements are made were identified; individual VLF comparison techniques used in the precise determination of each were examined. In part 2, the methods by which frequency standard oscillators are compared against precise (usually atomic) house standards such as the rubidium and cesium standards covered in part 1 are discussed, and multiple VLF techniques, typically used in navigation, which allow for time and frequency determinations to submicrosecond accuracies, are described.

comparing frequency standards

Since all aspects of determining time are usually based on the accuracy of an oscillator, let's see first how we can measure the absolute accuracy of oscillators. To be useful, any method for comparing oscillators against an atomic frequency standard must be able to resolve extremely small frequency differences. The following is an examination of five methods for doing this. The first two methods use an oscilloscope, while the last three methods use a period-measuring technique.

Lissajous patterns

Two frequencies can be compared by examining the Lissajous pattern that results from their introduction on an oscilloscope's horizontal and vertical inputs, respectively **(fig. 1)**. If the two frequencies form an integer ratio (a whole number), then the pattern will consist of a number of stationary loops in direct proportion to this ratio (see **fig. 1**).

If the two frequencies are very close to one another, an ellipse results. Slight frequency differences cause this ellipse to roll repeatedly through all orientations from 0° to 360°. If this duration (time of one "roll") is timed, then the difference in frequency is equal to the reciprocal of this time. Consequently, to match the frequency of any oscillator to that of a known standard adjust the oscillator until the ellipse is stationary. This method can be used to provide resolution approaching 1 part in 10^9.

oscilloscope pattern drift

An oscillator can be compared to a frequency standard by *externally triggering the* oscilloscope *from* the standard while a pattern of several cycles of the oscillator output is displayed. (**Fig. 2** illustrates this technique.) The ratio of the drift displayed on the oscilloscope is directly proportional to the error frequency of the oscillator under test. Using **fig. 2** as a guide, assume that an oscilloscope is used to check the time base in an HP5345A frequency counter against a time and frequency standard such as the HP5065A[1] Rubidium Vapor Frequency Standard.

If it takes 100 seconds for the pattern of two 10 MHz signals, to drift one cycle, then the error is 1 part in 10^9. Since:

$$\frac{\Delta f}{f} = \frac{0.01}{10^7} = 1 \times 10^{-9}$$

If two 100 kHz signals are being compared and the

By Vaughn D. Martin, 114 Lost Meadows, Cibolo, Texas 78108

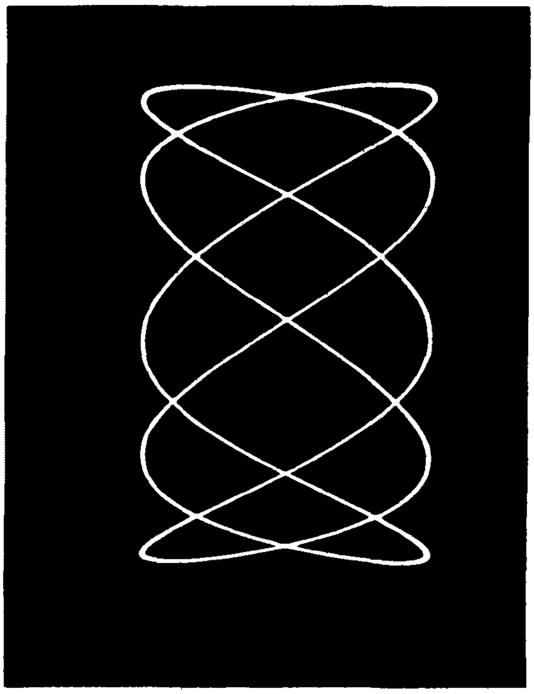


fig. 1. Lissajous oscilloscope pattern.

width of one cycle takes one second, then the frequency error is:

$$\frac{\Delta f}{f} = \frac{1}{100 \times 10^3} = 1 \times 10^{-5}$$

direct frequency comparison with a counter

You may have been wondering, as I used to, how a frequency counter operating on the principle of counting a set number of events during a fixed interval of time (typically one second), could measure a standard crystal oscillator and resolve any minute inaccuracies to a fine degree. After all, a typical frequency counter's timebase itself is far less accurate than the unknown standard crystal oscillator being tested. This "constructive skepticism" goes to the very heart of the issue in that the frequency counter is merely an instrument that takes the difference of two frequencies by having its own internal frequency standard substituted or bypassed by a standard of known accuracy.

As recent as the late 1960's the industry standard frequency was 1 MHz; it is now 10 MHz, with 1 MHz output still available as an option. (This division by ten is easily accomplished by applying the 10 MHz signal to a divide-by-10 IC such as a 74LS90.)

This method substitutes the reference oscillator for the frequency counter's time base to establish the sampling interval or period. Several cycles from the unknown oscillator are counted during this very precisely defined interval. Assume we are measuring a 1 MHz crystal oscillator against a known 1 MHz crystal oscillator standard. The counter's inherent $\pm$ count ambiguity limits the precision to 1 part in 10^6 for a one-second interval or sampling period. A ten-second interval or sampling period increases the accuracy to $\pm$ 1 count in 10^7. Let us assume we want to resolve the frequency to a very fine degree and have $\pm$ 1 count in 10^{11} as our goal. This would take a sampling period of 27.78 hours or 100,000 seconds.

Since this is not practical, another method has been devised to speed this process. The oscillator under test is frequency translated by a factor of 1,000 to a 1 GHz rate. (This can be done with a phase-locked loop.) What results is not only an increase in the fundamental frequency of 1 MHz to 1 GHz but, also an increase in the errors or deviations from 1 MHz by a factor of 1,000. Magnifying these errors by 1,000 also decreases the sampling period required by a factor of 1,000.

Some counters incorporate electronic counter-heterodyne frequency converters which enable the determination of frequency offset, relative aging characteristics, and fractional frequency comparison of deviations (if the measurement interval is kept below 10 seconds).

In modern frequency counters such as the HP5345A time base substitution by an extremely accurate standard can be effected. The same technique used to determine the crystal's accuracy as outlined in the previous section is used. Error is introduced by:

1) least significant digit (LSD) count

2) time base inaccuracy (the quality of the standard used)

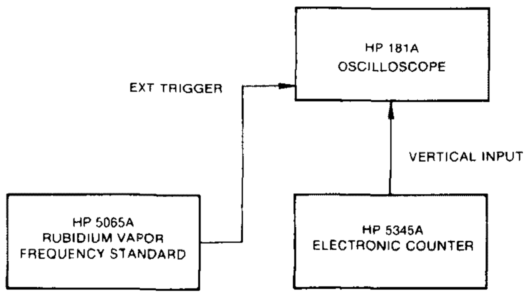


fig. 2. Oscilloscope pattern drift determination.

3) trigger error (start of sampling period)

Trigger error for this specific counter is less than 0.3% divided by the number of periods averaged for a 10 mV input sinewave, with noise specified at 40 dB below the signal level. The setup is shown in **fig. 3.** Let's go through a typical measurement and see how the calculations are done. Returning to our basic equation for error sources:

error max $= \pm$ *count* $\pm$ *trigger error* $\pm$ *oscillator errors*

Worst-case conditions for accuracy determination must assume errors as being all in the same direction and cumulative, although in actuality some errors could partially cancel or offset one another.

$$Error\ Max = \pm\ 2\ counts \pm \frac{0.3\%}{5 \times 10^{6*}} \pm 1 \times 10^{-11}$$

*5 MHz cesium beam standard

$= \pm\ 4.61 \times 10^{-9}$

for a 10 second gate time

$= \pm\ 4.7 \times 10^{-10}$

for a 1,000 second gate time

$= \pm\ 0.46 \times 10^{-11}$

Note how the accuracy is improved or the maximum error minimized as the sampling period increases, as previously discussed.

frequency comparison with a vector voltmeter

This is the last of the five methods of frequency comparison. A phase-sensitive or vector voltmeter is a special type of AC voltmeter with a high input impedance which minimizes the effects of circuit loading. It possesses some special features, including a filter to suppress harmonics and noise from the input source and a phase-sensitive demodulator. This circuit gives the instrument the ability to accept two input signals and determine how closely the unknown input signal's frequency overlaps or coincides in time with a frequency standard used as a reference input.

If the two signals are exactly in phase and at the same frequency, the phase angle generated is zero. As the frequencies depart from one another, the phase angle increases. In our particular case, the two frequencies, that of the standard or reference and that of the unknown frequency being measured, are so close that the angle will be very small. Therefore, a vector voltmeter is used that produces an output voltage in direct proportion to the meter's reading. This small voltage, which usually comes out the instrument's rear panel through a jack, can be amplified and a hard copy recorded on a chart recorder. This method of measurement can resolve differences in frequency to 1 part in 10^{13} in a matter of several minutes. A typical calculation will illustrate this technique:

$$\frac{\Delta f}{f} = \frac{\Delta\theta}{360\, t \cdot f}$$

Where:

Δf = frequency difference between the two signal sources (Hz).

f = frequency of the standard source (Hz).

$\Delta\theta$ = phase change in degrees during the measurement time.

t = time, in seconds, during which $\Delta\theta$ was measured.

$\frac{\Delta f}{f}$ = fractional offset of the source being checked.

If our standard and source to be tested both have 10 MHz outputs, and the angle indication of $\Delta\theta$ is 2.5° and the time is 120 seconds, then:

$$\frac{\Delta f}{f} = \frac{2.5^\circ}{360^\circ\,(120\ S)(10 \cdot 10^6\ Hz)}$$

$$= 5.78\ parts\ in\ 10^{12}$$

time comparison systems

Maintaining a consistent local system of time and frequency standards requires that they be compared against one another with a reference being established and maintained. The transfer of time information from one location to another is referred to as "time synchronization," whereas the transfer of data concerning frequency is referred to as "calibration." **Table 1** lists some of the more common methods of accomplishing the transfer of time and frequency data.

Radio broadcasts from frequency and time standard stations are the method most often used to establish the link that maintains this reference. The following is a discussion of both HF and LF/VLF techniques for maintaining local frequency standards

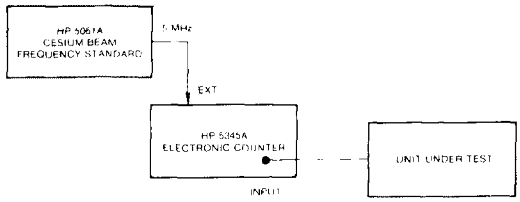


fig. 3. Direct frequency comparison with a counter.

* See **fig. 3**

table 1. A comparison of time and frequency transfer techniques.

TRANSFER TECHNIQUES		STATUS (1)	ACCURACY-FREQUENCY SYNCHRONIZATION	ACCURACY FOR DATE TRANSFER	AMBIGUITY (4)	COVERAGE FOR STATED ACCURACY	% OF TIME AVAILABLE	RELIABILITY	RECEIVER COST FOR STATED ACCURACY	COST PER CALIBRATION	NUMBER OF USERS THAT CAN BE SERVED	OPERATOR SKILL REQUIRED FOR STATED ACCURACY
VLF RADIO	COMMUNICATION/SFB GBR NBA WWVL	O	1×10^{-11} (3)	ENVELOPE 500 μs	PHASE ~50 μs	GLOBAL						
	NAVIGATION SYSTEM OMEGA	O/P	$<1\times10^{-11}$ (3)	≤ 10 μs■	PROPOSED CODE 1 YR / PHASE ~100 μs	GLOBAL			MODERATE			TIME CODE
LF RADIO	STANDARD FREQ. BROADCAST (WWVB)	O	1×10^{-11} (PHASE 24h) (3)	ENVELOPE ~1 ms	1 YR	USA (WWVB) LIMITED			MODERATE		USA (WWVB) EUROPE OTHERS	
	NAVIGATION SYSTEM LORAN C	O	1×10^{-12} GND (3)	~1 μs (GND) 50 μs (SKY)	50 ms / PHASE 10 μs	SPECIAL● AREAS					SPECIAL AREAS	
HF/MF RADIO	STANDARD FREQ. BROADCASTS (WWV)	O	1×10^{-7}	1000 μs	1 DAY / 0.5 min	HEMISPHERE		DEPENDS ON CONDITIONS				
	NAVIGATION SYSTEM LORAN A	O	5×10^{-11} (2)	2.5 μs NOT UTC		LIMITED AREAS		DEPENDS ON CONDITIONS			SPECIAL AREAS	
TELEVISION (VHF/SHF RADIO)	PASSIVE LINE 10	O	1×10^{-11} (24h)	~1 μs	1 DAY / ~33 ms	NETWORK COVERAGE	"LIVE" PROGRAMS				USA FOR EXAMPLE	
	ACTIVE LINE 1 (NBS TV TIME SYSTEM)	E	1×10^{-11} (< 30 min)	< 100 ns▲	1 DAY / ~33 ms	NETWORK COVERAGE					USA FOR EXAMPLE	
SATELLITES (VHF/UHF/SHF RADIO)	STATIONARY SATELLITES (TRANSPONDER) ONE WAY	E/O	1×10^{-10} (24h)	10-50 μs	DEPENDS ON FORMAT	HEMISPHERE	STATIONARY					
	STATIONARY SATELLITES (TRANSPONDER) TWO WAY	E/O	1×10^{-12} (24h)	~100 ns	DEPENDS ON FORMAT	HEMISPHERE			MODERATE			
	ON BOARD CLOCK (ACTIVE) ONE WAY - LOW ALTITUDE	O	$\sim1\times10^{-10}$ (24h)	0.5-50 μs	DEPENDS ON FORMAT	WORLD	10-15 min PER PASS 2-4 PER DAY	CLOCK NEEDS ADJUSTMENT				
SHF RADIO	MICROWAVE	E/O	$\sim1\times10^{-11}$ (PER WEEK)	≤ 100 ns	PHASE COMPARISON	LOCAL LINKS						
	VLBI	P	5×10^{-14}	~1 ns	DEPENDS ON FORMAT	HEMISPHERE	AS NEEDED					
PORTABLE CLOCKS	PHYSICAL TRANSFER	O	1×10^{-13}	100 ns●	1 DAY	LIMITED BY TRANSPORTATION	AS NEEDED		NONE			
	AIRCRAFT FLYOVER 2 WAY	E	1×10^{-12}	≤ 100 ns	DEPENDS ON FORMAT	LIMITED BY TRANSPORTATION	AS NEEDED					
PULSARS	OPTICAL SIGNAL — NP 0532	P	1×10^{-12}	~10 μs	~33 ms	HEMISPHERE	NIGHTTIME					
AC POWER LINE	POWER NETWORK SYSTEM	P	1×10^{-8}	~1 ms	16.7 ms	CONTINENTAL USA			MINIMAL	MINIMAL	CONTINENTAL USA	

GOOD ☐ FAIR ☐ POOR ■

NOTES: (1) Status of technique indicated as follows: O—Operational; P—Proposed; E/O—Experimental operational. (2) Estimates of day-to-day measurements within 2000 km (1250 mi) of Loran-A stations. These emissions not coordinated with UTC and manually operated crystal clocks drift. (3) From day to day phase measurements e.g., 1 μs per day phase change approximates 1 pt. in 10^{11} in frequency difference. (4) Left-hand designation gives the shortest time interval that cannot be resolved. Right hand number gives basic ambiguity. ●, by ground wave 1600 km; by sky wave thousands of kilometers depending upon conditions. ■, with proposed time code. ●, closure after 1 day. ▲, within local service area of TV transmitter and path delay known.

traceable to the NBS (National Bureau of Standards).

HF radio reception

HF radio transmission links suffer from propagation delays that are difficult to determine to better than 1 millisecond because of the changing ionosphere which is continually affected by sunspot, diurnal and seasonal variations. In addition, the number of propagational hops is also difficult to determine for distances greater than 2135 miles (3500 km).

Therefore, to use HF timing signals most effectively in the tick phasing adjustment method, to be explained next, these four recommendations should be followed:

- Select the highest frequency that provides consistent reception.
- Make all measurements in the day or night; avoid twilight or transitional hours.
- Observe tick transmissions for a few minutes to judge propagation conditions. Signals must exhibit minimum jitter and fade.
- Make time comparisons using ticks with the earliest arrival time (shortest propagation distance mode).

time comparison by tick phasing adjustment

This technique (**fig. 4**) compares local time against

time signals transmitted by an HF standards station such as WWV. The local frequency standard, typically a rubidium standard with a 10 MHz output, has this signal divided down to a frequency of 100 kHz and applied to a frequency divider and clock. This instrument further divides this signal to produce a 1 PPS pulse. These pulses are applied to the external sync of an oscilloscope while the vertical input to the scope is driven by an HF receiver that supplies the WWV ticks. This HF receiver is usually a good communications-quality instrument which can be tuned to the required frequencies (2.5, 5.0, 10, 15, and 20 MHz for WWV). It is preferable to use an antenna that is directional and is oriented in a manner favorable to the transmission mode consistently providing the shortest propagation path.

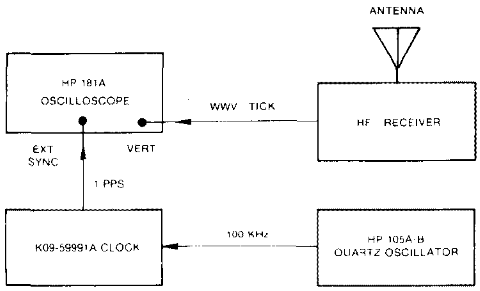


fig. 4. Time comparison by "tick phasing" adjustment.

The clock incorporates time delay thumbwheel switches to control the start of the oscilloscope sweep by delaying it until the WWV "tick" begins the sweep. By successive adjustments of this control, the two ticks are brought to near-coincidence. The WWV tick is a 5 ms pulse of 1 kHz energy.

The operator must be skillful at determining when the tick arrives. A tick-averaging scheme helps in this regard. An oscilloscope with a variable persistence screen permits the operator to view repeated sweeps of WWV displayed together (see **fig. 5**). An alternative tick-averaging method is to make an oscillogram using a scope camera. With either method, time comparison readings are possible once the scope's sweep time has been properly calibrated.

LF and VLF compared to HF signals

Variations in the ionosphere ion and electron density versus height cause propagation time of a signal to change continually, refer to **fig. 6**. This necessitates that the data recorded from stations such as WWV be averaged for days to remove anomalies in the data in order to obtain at best 1 part in 10^8 accuracy — note that WWV signals are stable to 5 parts in 10^{11} as transmitted.

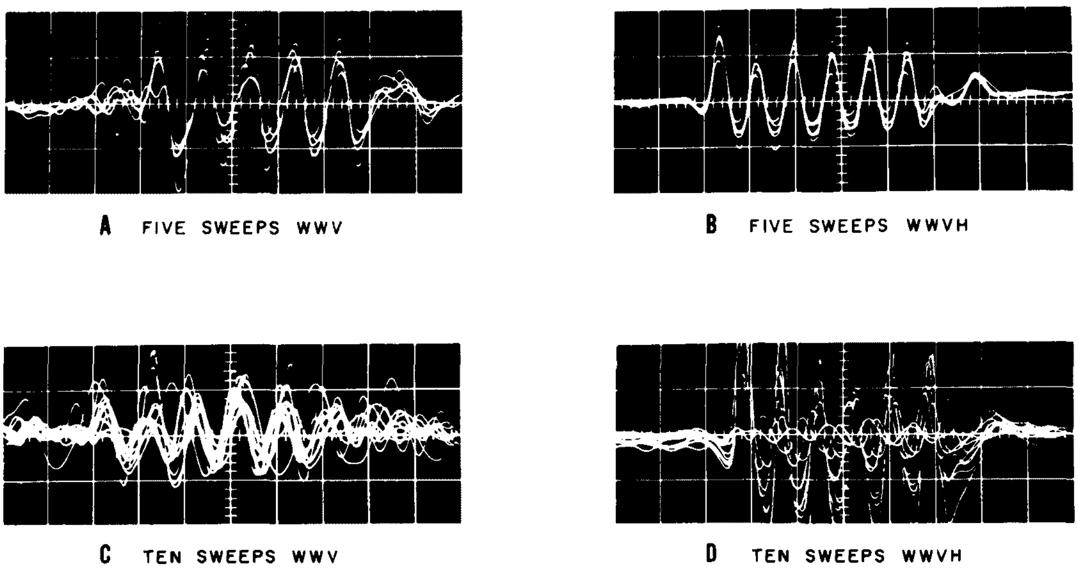


fig. 5. Photographic tick averaging adjustment.

Low frequency and VLF signals are far more stable than HF signals. This is because, unlike HF signals, they follow the curvature of the earth and follow a "duct" created by the earth's surface and the ionosphere. Since the earth's ionosphere acts as a boundary rather than as a reflector, as it does with HF signals, its variations have a much less pronounced effect. Therefore, for standard frequency transmissions this more phase stable method is used. When precise timekeeping is of paramount importance, most users still rely though upon the HF service provided by WWV. This is because of two factors: First, information bandwidth characteristics have limited to some extent the use of lower frequency signals for time-of-day and time comparison information. Secondly, to realize very high precision in clock synchronization, a fast risetime signal for a time marker is required; unfortunately the LF/VLF antennas used have large time constants which prevent this from being possible or practically realizable.

LORAN-C using cesium beam standards

Now that we have a better appreciation of the characteristics of sky and groundwaves, let's examine a navigation system that is stabilized by cesium beam standards and is one of the most accurate time transfer media available through radio waves.

LORAN-C (*LO*ng *RA*nge *N*avigation) is a navigation system used in some parts of the Northern Hemisphere as a method of time and frequency distribution. Operated by the U.S. Coast Guard, this navigation system uses pulsed transmissions on a carrier frequency of 100 kHz with a 20 kHz bandwidth.

The LORAN-C system has the following advantages and disadvantages, respectively:

1) The transmitters are controlled by cesium standards.

2) Propagation delays of groundwaves still provides ± 0.3 microsecond accuracies.

3) The TOC (time of coincidence), to be explained shortly, is provided in advance by the USNO (United States Naval Observatory).

4) Equipment costs are reasonable.

The disadvantages are:

1) Mixed terrain (sea and land) result in hard-to-predict effects of the propagation delays of skywaves.

2) Local clock time must be very exactly known, as will be shown shortly.

3) Coverage is not yet global.

4) Cycle selection requires a skilled operator.

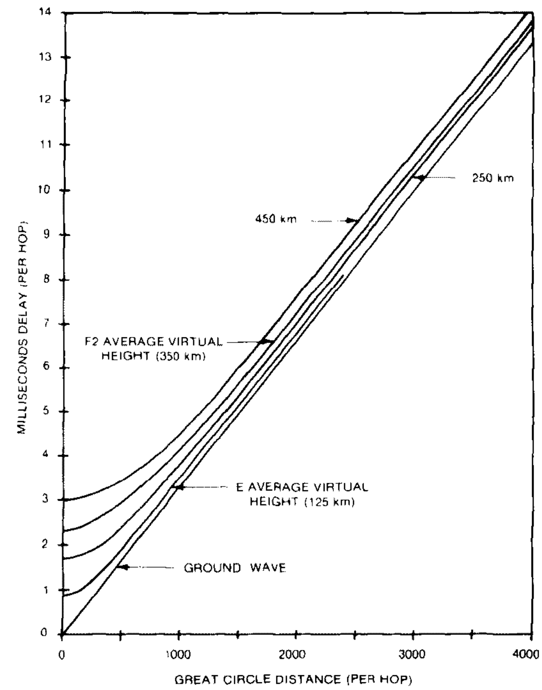


fig. 6. Transmission delay experienced on groundwave, E and F2 layer propagation modes.

time and frequency determination using LORAN-C

This navigation system *does not* broadcast a time code signal. It is therefore necessary to know the time-of-coincidence (TOC) of a LORAN-C signal relative to a UTC (universal coordinated time) second. Each LORAN-C chain transmits a unique format. Within the GRP (Group Recognition Period), which is unique to each chain, the master station transmits precisely spaced groups of nine pulses (**fig. 7**). Each of the slaves transmits eight pulses within the GRP.

There is only one GRP that will provide a TOC every second. And the period between a pulse coinciding with one UTC second and another pulse-UTC second is a function of the repetition rate of the whole chain. Each chain is assigned a different repetition rate and this and other data typical of this chain can be found in USNO documents.

Figure 8 is a typical system to utilize LORAN-C transmissions for time transfer. This system has a 1 PPS signal used to start a time interval counter with the output of the LORAN-C receiver used to stop the interval counter. The output of the LORAN-C receiver is a 1 PPS signal phaselocked to the received signal and in sync with the TOC. By using the count

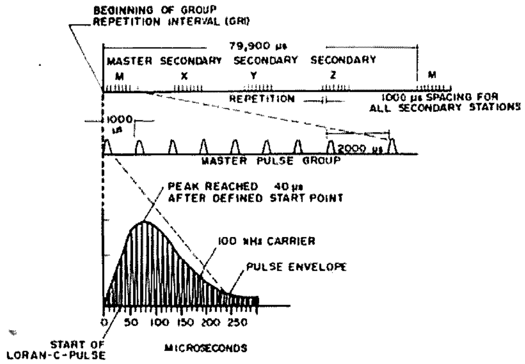


fig. 7. LORAN-C pulse group formats.

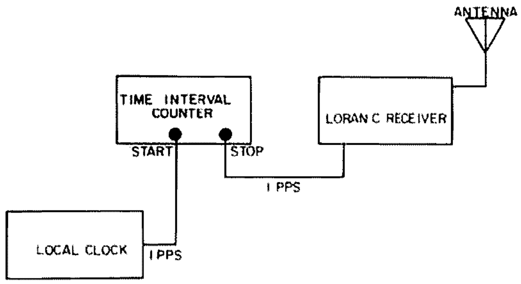


fig. 8. A typical LORAN-C timekeeping system.

on the interval counter in conjunction with USNO published data, you can determine the difference between UTC, USNO, and the local clock.

multiple VLF systems

Single VLF systems, with their less precise clock synchronizations, are unacceptable in highly demanding applications such as LORAN-C and Omega navigation systems. What is required in these instances is submicrosecond accuracies, obtainable only with multiple-frequency VLF techniques which use two or more closely spaced, coherently related, sequentially transmitted signals.

The multiple-carrier VLF techniques function by extracting timing data in the difference frequencies of two carriers. This allows for individual cycle identification of one of the carrier frequencies. Used successfully with WWVL and Omega frequencies, this technique requires signals, transmission media and receiving/comparison equipment that are extremely stable. The system functions by setting coarse time using an HF radio transmission (link) having an accuracy of several milliseconds which serves to resolve any initial frequency ambiguities. The multiple carrier VLF method includes a local calibration signal for simulating the frequency of the received signal to relate the local time scale to that of the transmitter. Agreement between the received and calibrated VLF phases is made systematically, and the local clock phase-shifter adjusted, until all simulated signal phases are identical to the actual received signal phases for a single setting of the phase shifter. This phase relationship remains essentially unchanged (except for clock interruption and phase loss), and the VLF receiver can be turned off and on without affecting the calibration.

The development of precise time and frequency standards, and the design of equipment to perform extremely accurate tests and measurements, is one of the most promising areas in the field of communications. As satellite communications technology advances, an equivalent degree of progress should be evident in this specialized field.

references

1. Vaughn Martin, "Time and Frequency Standards: Part 1," *ham radio*, November, 1983, page 36.

ham radio

when hazardous waste comes home — PCBs in the ham shack

What's inside your dummy load? Knowing the answer may be important.

So far we at *ham radio* have been lucky: hazardous waste has usually been a matter of concern for somebody else. Because we didn't live near the Love Canal, or along the 210-mile stretch of North Carolina highway poisoned by a "moonlight dumper," there seemed to be time and room for complacency.

But within recent months, two public wells near ham radio's office — one fifteen miles east, the other fifteen miles west — were added to the EPA list of over 500 contaminated sites eligible to compete for $1.6 billion in "Superfund" money. Even closer to home, reports in the Amateur press have pinpointed dummy loads as a potential source of PCB contamination in the hamshack.

CDC warning

Katherine Hevener, WB8TDA, called attention to the possible hazard in the May 15 issue of *New England Report/Crossbander:*

...According to CDC Public Health Advisor David Forney, the potential danger from the dummy loads could come from a type of transformer oil containing polychlorinated biphenyl (PCB), a substance known to cause liver damage in humans and animals. Fifteen samples of transformer oil will be taken from dummy loads in Houston, Atlanta, St. Louis, Denver, Columbus (Ohio), and Boston. ...Although hams should not panic, Forney urges those concerned to take precautionary measures, such as using dummy loads in well-ventilated rooms, and preventing the possible PCB-contaminated transformer oil from vaporizing.[1]

When we called Forney for an update in early September, he was careful to note that these precautions apply even to *non-leaking* dummy loads — loads without visible defects and functioning without apparent cause for concern — which could release invisible airborne PCBs when they overheat. Consequently, it's important to give your dummy load a rest. *Don't let it overheat*: inhaled PCBs may be as dangerous as PCBs that are ingested or absorbed through the skin. *Follow the rating curve* (**fig. 1**) *for your dummy load;* one should be found in your copy of the owner's manual for the model you have. If you don't *have a rating curve*, get a copy from the manufacturer.

According to Forney, *leaking dummy loads require immediate attention.* A leaking canister — regardless of however slight the leak — suggests the immediate possibility of direct absorption through the skin, regardless of whether the dummy load is in use or not.

Don't risk handling a leaking dummy load until you know what's in it.

what's inside?

If your dummy load was sold with oil already in it, contact the manufacturer to determine whether or not it was filled with transformer oil containing PCBs. Obviously, if you built your dummy load from a kit and installed your own oil, you know whether it contains mineral oil or transformer oil. If it's mineral oil, there's no cause for concern beyond the knowledge

By Dorothy Rosa Leeds, assistant editor, *ham radio*

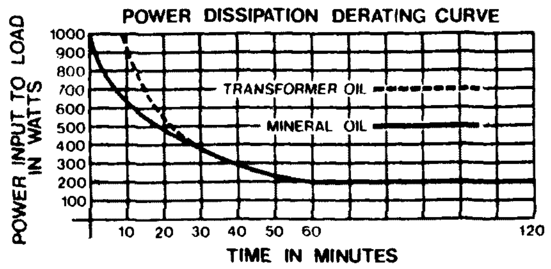


fig. 1. Rating curves for Heathkit Cantenna (Model HN-31A) provide data for dummy load in mineral oil as well as in transformer oil. Since specifications differ, product literature or manufacturer should be consulted for capacity of other dummy loads. (Note: Transformer oils made without PCBs are available: for example, Dow Corning manufactures a silicone-based liquid said to offer excellent dielectric properties, yet be environmentally safe.)

that mineral oil will burn at a comparatively low temperature. PCB oil will not; in fact, it was its excellent thermal and chemical stability, as well as its superior dielectric and insulation qualities that made PCB oil such a resounding commercial success from its introduction in the late 1920s through its decline in the late 1970s.

If your kit-built dummy load contains transformer oil, then identifying its composition may be more difficult. Testing for PCBs, down to the parts-per-million level, can be done in any laboratory with a gas chromatograph and a mass spectrometer, but this testing can be expensive — more than the cost of a new air-cooled or silicone oil-cooled dummy load.

biomagnification

The same qualities that made PCBs a boon to industry have made them an environmental scourge: they simply don't break down into new chemical arrangements. According to a U.S. Navy report, PCBs "biomagnify in the food chain — that is, they accumulate in the tissues of living organisms and as one organism feeds on another, progressively greater concentrations occur as the food chain progresses upward towards man."[2] Current estimates are that 99 percent of Americans carry a measurable amount of PCBs in their fatty tissues.[3] But unfortunately, the human body has no way to break the compound down into its less harmful components.

In structure, PCBs are similar to DDT (see **fig. 2**). Once contaminated by DDT, the human body requires up to fifteen years to rid itself of the chemical, but this can happen only if no additional DDT is absorbed. The federal ban on DDT in 1972 has effectively controlled this hazard, but despite their structural similarity to DDT, PCBs are significantly more dangerous, more prevalent and much more difficult to eliminate both from the body and the environment.

advice from the experts

Throughout their nearly 50 years of manufacture, PCBs were used primarily in electrical transformers — including the familiar "pole pig" sometimes acquired for use as a power supply — and capacitors, heat transfer systems, and hydraulic systems. They were also extensively used as plasticizers in paints, inks, adhesives, sealants, caulking compounds, coatings, and in carbonless copy paper, as well as in fluorescent lamp fixture ballasts and in pesticides and weed-control products.[4]

Suppose you've concluded that your dummy load contains, or may contain, PCB's. What then?

Don't put it out with the trash.

Al Hyer, of the EPA public affairs office in Washington, D.C., told us that the individual owner of any suspect device is solely responsible for its disposal, and urged that concerned Amateurs get professional advice before disposing of any questionable dummy load. He suggested calling your state environmental agency first; if an appropriate response and timely

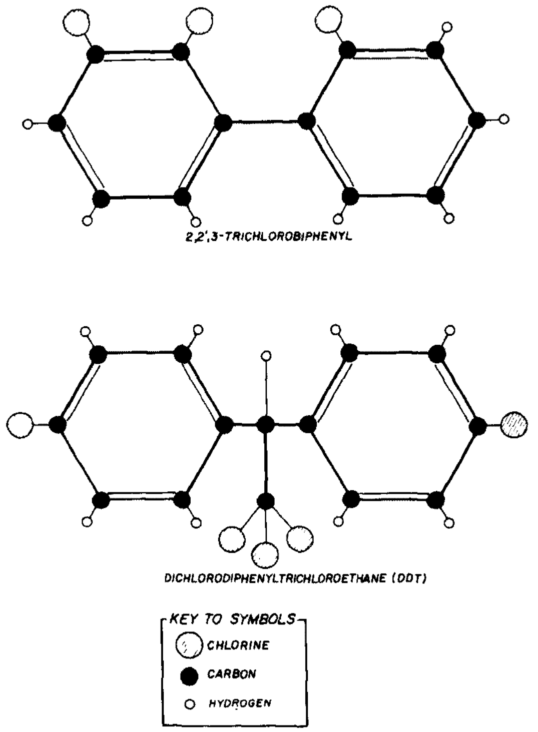


fig. 2. 2, 2', 3 trichlorobiphenyl is one of several PCBs, all resembling DDT in chemical structure.

assistance are not immediately forthcoming, then the regional office of the EPA should be called (see **table 1**). "Help is available," Hyer said, "and it may not cost a dime."

Another knowledgeable source we contacted was less optimistic about obtaining help from the public sector. Alan Borner, Executive Director of the Environmental Hazards Management Institute, an independent, non-profit educational and consulting firm in Portsmouth, New Hampshire, cited the potential difficulty of calling the attention of a large bureaucratic state agency to a seemingly minor — yet perhaps dangerous — problem. Borner suggested that Amateurs turn to utilities or industries instead of state or federal agencies. "Call the environmental manager for the local public service company," he said. "Try any major chemical corporation or utility. Ask for the national environmental program manager and get his or her advice. These people have been walking through this morass for years."

Public utilities are especially sensitive to PCB questions, Borner noted, because under federal law, owners of potentially hazardous materials are responsible for them. Even though vandals still shoot out transformers for sport, it's the utilities who foot the bill for replacements and any lawsuits that may result.

According to Borner, the number and availability of disposal facilities for PCBs is extremely limited, so disposal, however it is ultimately accomplished, may be complicated by red tape and delay. But if only because PCBs are the only compound specifically singled out for immediate restriction as hazardous in the Toxic Substances Control Act of 1976, their handling and ultimate disposal should be taken seriously.

Another option open to Amateurs seeking to dispose of questionable or contaminated dummy loads is to contact an independent hazardous waste contractor. Some firms will deal only with industrial clients and have minimum quantities they will accept. A few large firms — SCA Chemical Services was one we spoke with — have limited programs designed to help individuals dispose of hazardous materials in a responsible manner. (SCA sponsored a "household cleanup" program on Cape Cod last fall, and other efforts are planned in certain areas of the Northeast.)

There are also smaller firms that will accommodate small quantities of waste as well as industrial quantities. Transformer Service, Inc. (TSI), of Concord, New Hampshire, for example, offers testing and disposal services. Louis LaSalle, TSI's Northeast Sales Manager, told us his firm would assist Amateurs in disposing of contaminated dummy loads. Group rates are negotiable for both testing and disposal services.

table 1. List of administrators designated as "hazardous waste contact persons" at regional offices of the Environmental Protection Agency (as of July 29, 1983).

Region I — Ira Leighton, Environmental Protection Agency, John F. Kennedy Building, Boston, Massachusetts 02203, (617) 223-3468

Region II — Dr. Ernest Regna, Environmental Protection Agency, 26 Federal Plaza, New York, New York 10278, (212) 264-0504/5

Region III — Anthony Donatoni, Environmental Protection Agency, 6th and Walnut Streets, Philadelphia, Pennsylvania 19106, (215) 597-7937

Region IV — Allan Autley, Environmental Protection Agency, 345 Courtland Street N.E., Atlanta, Georgia 30308, (404) 881-3016

Region V — Karl J. Klepitsch, Jr., Environmental Protection Agency, 230 South Dearborn Street, Chicago, Illinois 60604, (312) 886-7435

Region VI — Ms. Pat Hull, Environmental Protection Agency, 1201 Elm Street, First International Building, Dallas, Texas 75207, (214) 767-9736

Region VII — Chet McLaughlin, Environmental Protection Agency, 324 E. 11th Street, Kansas City, Missouri 64106, (816) 374-6534

Region VIII — Jon P. Yeagley, Environmental Protection Agency, 1860 Lincoln Street, Denver, Colorado 80295, (303) 837-2221

Region IX — Laura Yoshii, Environmental Protection Agency, 215 Fremont Street, San Francisco, California 94105, (415) 974-8127

Region X — Tobias A. Hegdahl, Environmental Protection Agency, 1200 6th Avenue, Seattle, Washington 98101, (206) 442-2808

Note: *The EPA operates a special hotline dedicated specifically to questions about PCBs: 1-800-424-9065. Questions about other hazardous wastes should be directed to another EPA hot line: 1-800-424-9346.*

LaSalle cautioned Amateurs to choose their disposal contractor with care, and to insist on receiving documentation confirming appropriate incineration of the waste oil. He confirmed Hyer's and Borner's warning that the owner is responsible for safe disposal: "You can't contract your liability away...anytime the material leaves your premises, you've lost control; be sure you know where it goes," he warned.

don't inhale

According to LaSalle, dummy loads can be shipped; check with your contractor for specific instruction for safe handling and packaging.

In a letter to *ham radio*,[6] Tom Runyon, VE5UK, took the media to task for "overrating" the potential

hazards of PCBs, but at the same time expressed appreciation that the question had been brought to the attention of Amateurs as a group. He gave this advice for distinguishing PCB oil from mineral oil in dummy loads:

1. *The smell of PCBs is somewhat similar to that of moth balls. Ordinary vegetable or mineral trans-former oils smell like "oil."*

2. *Pure PCB is heavier than water, and a drop dropped into a bottle of water will sink. Ordinary transformer oil will float on water.*

(We don't recommend that you actually sniff the oil in your dummy load; inhalation, as noted before, may be hazardous. And before you follow any of VE5UK's advice, check with your regional EPA office or other appropriate authority.)

Runyon pointed out what may be an important distinction in any review of this issue: that there's a difference between acute poisoning with PCB's in high concentrations and short or long-term exposure to materials containing PCB's in lesser concentrations. Serious health effects have been documented in episodes involving heavy concentrations of PCB's, but to our knowledge, there is no evidence at this point to confirm the common fear that incidental exposure to low levels of PCBs has any documentable health effects. *We just don't know. Nobody knows.*

Hearing that nobody really knows for sure about the effects — if any — of low-level exposure to PCBs usually causes one of two reactions: either people assume that since nobody knows, there's nothing to be afraid of, or that *because* nobody knows, there's reason to be gravely concerned.

toward solving the problem

Congress was sufficiently concerned about the possibility of serious PCB-induced health hazards and passed The Resource Conservation and Recovery Act of 1976, in which the manufacture of PCBs or products containing PCBs, and the importation of any polychlorinated biphenyls, were declared illegal. No longer could industries or utilities drain the oil from a malfunctioning transformer, perform a repair, and replace the oil. It had to be discarded. *Where and how* became an entirely new problem.

With increased understanding of the importance of responsible disposal of hazardous wastes, industry has turned to two main means of disposal: incineration and detoxification by dehalogenation. (PCB waste in low concentrations may still be buried, but only in approved landfills under close EPA supervision. High-concentration PCB materials may not be buried.)

Incineration is effective only at temperatures in excess of 2300 degrees Fahrenheit; incomplete incineration at lower temperatures has been known to release dioxin (tetrachlorodibenzo-para-dioxin, TCDD), estimated to be approximately 500,000 times more dangerous than PCBs. (Its levels of contamination are measured in the parts-per-trillion, rather than in the more modest parts-per-million used to calculate the levels of PCBs.)

Several state-of-the-art incinerators are in operation now; the Rollins Environmental Services, Inc., facility in Deer Park, Texas, was the first to receive a permit from the EPA to incinerate PCBs in liquid and solid form. Rollins claims an efficiency level of 99.9999.[6] Several major chemical corporations have developed elaborate incineration systems (see **figs. 3, 4**), but the cost remains high — $9 to $12 per gallon, plus fifty cents per gallon for transportation alone.

Industries seeking to develop effective disposal systems face a costly and time-consuming challenge, however, because citizen resistance to incinerators remains intense. No wonder: even with high-efficiency incineration, the possibility of accident or unwanted emissions remains.

At least two experimental incinerators are currently operating at sea, with more in the planning stages, and Sunohio, Inc., has developed five mobile units that process PCB waste on-site.

Unfortunately, we read less about progress in the development of effective means of detoxification and effective solutions to the problem of disposal than we read about accidents involving PCB contamination of people, wildlife, and the environment. The stories *are* shocking. While research has yet to conclusively link inordinately high levels of cancer in certain areas with PCB contamination, studies have documented liver damage, digestive disturbances, jaundice, impotence, swollen limbs, and serious skin problems, including lesions, chloracne,* and pigmentation disturbances. The most notorious episode of PCB poisoning occurred in Japan in 1968, when 1,291 adults and children consumed rice oil contaminated with PCBs. Twenty-four died — nine of malignant neoplasms. In *Laying Waste: The Poisoning of America by Toxic Chemicals*, three-time Pulitzer Prize nominee Michael Brown, who covered the infamous Love Canal story in the *Niagara Gazette*, reports on a study of thirteen women who had ingested the contaminated rice oil during pregancy. Of the thirteen, "nine bore children with impaired liver function and other defects...."[7]

Japan outlawed the production of PCBs in 1972.

According to EPA estimates, there are 9,131 *abandoned* waste disposal sites. 440 million pounds of

*While chloracne is not life-threatening, it is disfiguring; its treatment — drainage and surgery — is said to be painful.

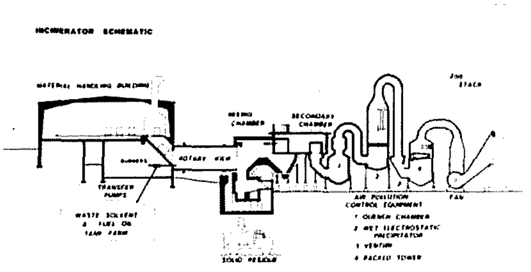


fig. 3. 3M Company's state-of-the-art incinerator at Cottage Grove, Minnesota, processes 500 55-gallon drums of hazardous waste daily. Constructed in 1971 at a cost of $4.6 million, and updated as technological improvements became available, the facility treats only 3M waste.

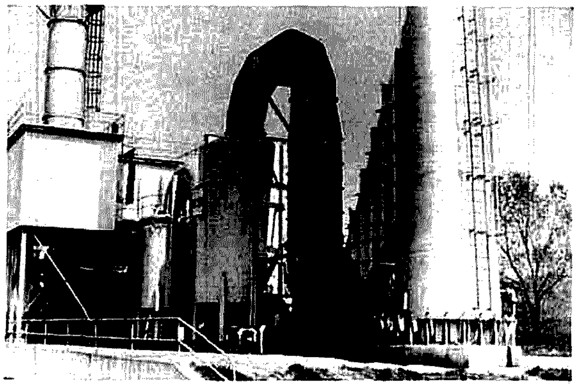


fig. 4. 3M incinerator air pollution control train includes a series of water showers designed to remove gas and particulate matter from air stream before it passes through 500-horsepower fan at base of stack, which measures 5 feet in diameter, 200 feet in height.

PCBs await disposal, and 750 million pounds remain in use or storage, with 40,000 of the 20 million transformers in use by utilities today containing Askarel, a fluid composed of 60 to 100% PCBs.[8] Of the 200 or more varieties of PCBs made, most can be identified by name; tradenames include Aroclor, Pydraul, Therminol, Pyroclor, Santotherm, Pyralene, Pyranol, Inerteen, Asbestol, Chlorextol, Diachlor, Dykanol, Elemex, Hyvol, NO-Flamol, Saf-T-Kuhl, Clorinol, Clorphen, and Eucarel.[8]

Because the number of approved incinerators is limited, and because the issue of incineration is controversial, industry has turned to detoxification as an alternative. In 1980, Goodyear Tire & Rubber announced the development of a method of breaking down PCBs into sodium chloride and sludge.

Only recently have the EPA, FDA, and Department of Agriculture ordered removal of equipment containing PCBs from food and feed preparation and storage areas. Currently under EPA study is a proposal to eliminate equipment containing PCBs from public buildings. (In 1976, 95 percent of all oil-filled capacitors manufactured in the United States were made with PCB oil.) PCBs are still very much with us.

By the time serious health implications were first suspected — in the early 1970s — the annual rate of PCB production in the United States had reached 34,000 tons (30,844 metric tons). But it was too late to reverse the worldwide spread; by 1970, traces of the compound had been found in almost every American river and even at the Arctic Circle, in the bodies of nonmigatory bears. Commercial and sport fishing, and duck hunting on many waterways has been seriously disrupted or destroyed by PCB contamination.

With the banning of PCB production in the United States, it would seem that at least one part of the problem has been solved. But the question of how to handle what we have already remains; the fact is that we as individuals, families, and as a nation, generate more household and industrial waste than we can effectively dispose of.

So the next time you sit down to operate, cast a wary eye on your dummy load. It may be a small piece of a very large puzzle.

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ham radio

CDC confirms PCBs in dummy loads

As this issue went to press, the CDC had reportedly found at least one PCB-contaminated dummy load in its study of samples taken from Amateur equipment in six cities (see page 42).

The contaminated dummy load belonged to Richard P. Beebe, K1PAD, ARRL section manager for eastern Massachusetts. Of the 15 samples taken from the Boston area, Beebe's was the only one found to contain PCBs.

Beebe told *ham radio* that he had been advised by the CDC to undergo a blood test in order to determine whether he had experienced any negative health effects as a result of exposure to the material.

According to CDC epidemiologist Paul Stehr, the CDC study will continue, as leads in which potential problems were identified are followed up and owners of questionable dummy loads notified.

Katherine Hevener, WB8TDA

photovoltaic cells: a progress report

Seminar covers PV basics, "breeder" concept, and Amateur Radio applications

In November, 1982, 400 people participated in what has been termed the "largest publicly held seminar" on photovoltaics in the world. The gathering, presented by ENCON and SOLAREX in Dearborn, Michigan, provided the latest information on this subject to a group consisting of Radio Amateurs, corporate engineers and building contractors. The following article discusses the important features presented at the seminar.

Early photovoltaic cells were manufactured solely for space applications, and between 1958 and 1973 cost $500 to produce one (peak) watt. At this rate, the early cells would have had to operate for 40 years to pay back the energy expended in their manufacture alone; in order to compete with conventional electrical power generation, the cost of the cells would have to decrease dramatically. By 1977 the cost of solar cells had dropped to $20 per peak watt, with a calculated energy payback of less than six and a half years. Today, costs are lower than $10 per peak watt, with an energy payback of less than three and a half years. Both the cost and energy payback are continuing to decrease.

fig. 1. "Breeder concept" structure uses solar energy to provide power for production of solar cells.

What this indicates is that the technology, productivity, and energy consumption problems are being resolved, and that the solutions to the remaining issues are in hand. It is believed that by the time photovoltaic demand rises to the 100 megawatt level, costs will drop below $4 per watt and the energy payback will be reduced to less than one year.

photovoltaic cell material

Historically, several materials have been used for experimentation in photovoltaic devices: gallium arsenide, cadmium sulphide, and silicon, among others. Silicon has been chosen for widespread use and is the logical material since it is the second most abundant element on earth, is low in cost when used in a relatively low purity form and provides high photovoltaic efficiency. One key to lowering costs of solar cells has been the development of semicrystalline silicon — a form of polycrystalline silicon developed specifically for photovoltaic applications.

Semicrystalline silicon is fabricated using a procedure known as Ubiquitious Crystallization Process (UCP), in which the silicon takes on the shape of the mold in which it is crystallized. One common configuration* is that of square wafers, which lends itself to high density packing in a solar panel. Perhaps most importantly, UCP uses inexpensive, low-purity raw silicon. This means that the expensive, highly pure and energy-demanding raw material used in the conventional single crystal process can be bypassed.

the breeder concept

One way to decrease the price of a product is to increase its productivity. One successful means of increasing productivity, called the breeder concept method, incorporates a large industrial structure completely powered by the sun. With its sloping, south-facing roof, the building features a 28,000 square foot (2600 square meter) array of SOLAREX semicrystalline silicon solar cell panels (see **fig. 1**).

By Paul J. DeNapoli, WD8AHO, ENCON Corp., 27600 Schoolcraft, Livonia, Michigan 48150

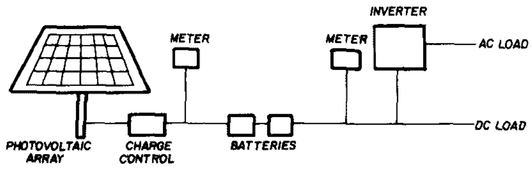


fig. 2. Block diagram of an electrical generation system based on a photovoltaic source.

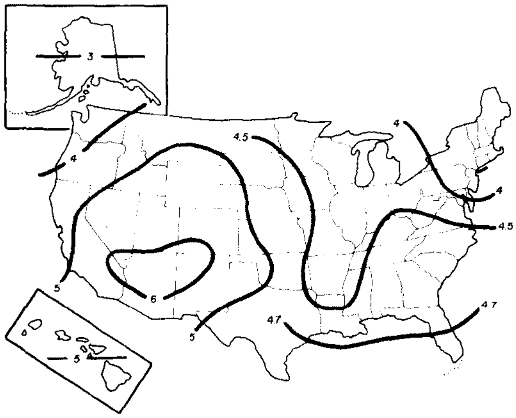


fig. 3. Average peak sun hours per day throughout the U.S. ("Guide to Solar Electricity," Solarex technical staff report, Solarex Corporation, 1335 Piccard, Rockville, Maryland 20850.)

The 200 kW rooftop photovoltaic array converts sunlight directly into electricity to provide power for the facility's production lines, which produce photovoltaic cells and panels. So the facility actually uses power provided by the solar cells on its roof to manufacture, or "breed," more solar cells. It derives all of its electric power, including power for lights, air conditioning, and production equipment, from the more than 3000 semicrystalline panels mounted on its roof. Its photovoltaic system, aided by solar thermal energy and thermal storage, also provides for all facility heating requirements, elminating the need for oil or gas heating. The 200 kW array produces an average of 800 kWh per day of energy coupled with a 2.5 MWh battery storage system. This storage capacity provides steady, uninterrupted power of 60 kW in the unlikely event of four dark days in succession.

powering an Amateur station

A complete system sufficient to power an Amateur Radio station under ordinary circumstances or during times of crisis would consist of photovoltaic panels, charge-controlling devices, and deep-cycle renewable energy batteries, (see **fig.** 2). A typical "barefoot" installation allowing for three hours of operation per day (one hour transmit, two hours receive) would require an average of four peak hours of sun in an area such as Michigan (see **fig. 3**), where ENCON and SOLAREX test facilities are maintained.

The transmitter load in this case would be 15 amperes for one hour or 15 ampere-hours. The receiver load would be 0.6 amperes for two hours or 1.2 ampere-hours. The combined load requirement would be 16.2 amperes, or more simply, 17 ampere-hours.

It's usual to factor in a system loss term of 20%, which raises our design requirements to a total of 20.4 ampere-hours. Referring to **fig. 3** and dividing 20.4 by 4 (average peak sun hours) provides our design goal of 5.1 amperes. Examining the industrial literature shows that the SX-100** PV panel produces 1.9 amperes at 17.0 volts or 32 watts under these conditions. Use of three of these panels meets our requirements *(1.9 × 3 = 5.7 amperes)* conservatively. **Fig. 4** has been included to assist readers in designing installations according to their own specific requirements.

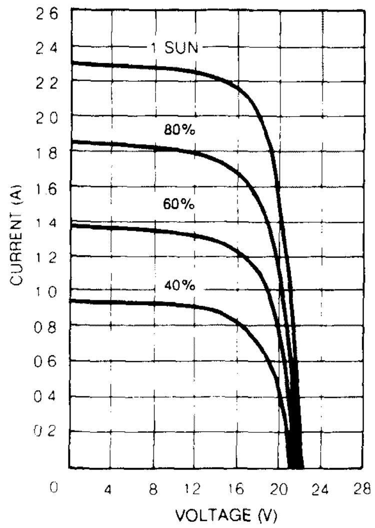


fig. 4. SX-100 solar panel performance at various light intensities, T = 25°C. (Courtesy of Solarex Corporation.)

*Manufactured by Semix Incorporated, a subsidiary of Solarex Corporation
**Manufactured by Solarex Corporation

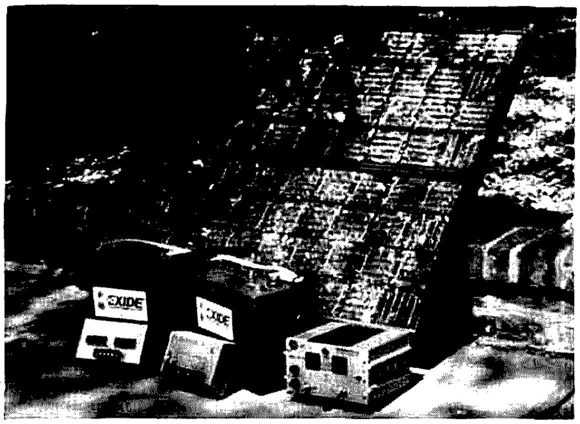


fig. 5. A typical "barefoot" ham station can be powered by a two-PV panel array with two 6 VDC 185 ampere-hour batteries as backup.

battery calculations

For battery (or storage) calculations, use a NO SUN day number for added reliability.

$$20.4 \text{ ampere-hour load} \times 7 \text{ (No SUN days)} = 143 \text{ ampere-hours}$$

$$143 \text{ ampere-hours} \times 2 \text{ (50\% battery drain)} = 286 \text{ ampere-hours}$$

Four 6 VDC 185 ampere-hour batteries will supply 370 ampere-hour capability at double the voltage or 12 volts DC (series-parallel configuration).

A list of materials for this system would include 3 SX-100 panels, 4 Exide 6 volt DC batteries, 1 charge controller, 1 mounting hardware kit, and 1 meter package and cable.

Due to SSB characteristics, a transmitter will never really see the full ampere output in transmitting. A system containing 2 SX-100 panels and 2 batteries with accessories can be utilized in a very energy-efficient manner for normal and emergency operations, (see **fig. 5**).

when will the price come down?

PV's have been dropping in price for twenty years. The pricing structure of photovoltaics is now at an affordable level, but without consumer support, we'll never see another decrease because of the elementary laws of supply and demand. A prime example of how consumer-supported products decrease in price are calculators and personal computers. The same principles apply to photovoltaics: until the Amateur Radio community chooses to support PV products, the industry will be slow to meet its ultimate goal of supplying reliable, cost-effective, primary and emergency power for communication, residential and commercial needs.

ham radio

vertical phased arrays: part 5

ABCD matrix parameters simplify network calculations

The most recent article of this series[1] discussed the application of lumped-constant circuits to the drive networks for all-driven element phased arrays. Design equations were presented for the most commonly used four-terminal networks. The design process and general procedures that must be followed for any drive network were reviewed, using typical arrays as examples.

Not discussed in part 4 was how the input/output calculations might be done. Those familiar with complex algebra and the use of a Smith chart can do these calculations one circuit branch at a time with these relatively simple networks. But it is a tedious process at best, prone to human error and cumulative errors resulting from rounding off in chain calculations. And one of Murphy's Laws asserts that errors are always committed at the *beginning* of the longest chain of calculations!

There is an alternative to this drudgery: matrix algebra. Using it allows us to determine the input conditions that result when any load is connected to a network. We do not have to calculate each circuit branch individually; all we need to know is the circuit type. There is even a built-in method of checking accuracy which aids in eliminating arithmetical errors and the entry of incorrect signs.

Matrix algebra has been used for solving problems of networks, transmission lines, and filters since the late 1940s because of its convenience in rapid circuit analysis and synthesis. With the advent of computers the use of matrices has increased, and it is possible that many readers have been using matrix methods without recognizing them as such, since the methods have often been incorporated into scientific computer programs as special process calls or library functions.

If matrix methods have been neglected in Amateur literature, it is not for lack of suitable applications. Perhaps the method has seemed too esoteric to be applied to such mundane problems as matching antennas to transmission lines, or that the jargon associated with it has frightened experimenters away; perhaps Amateurs have simply not been sufficiently exposed to this powerful mathematical tool.

four-terminal network matrices

Mathematical analyses using matrices require only the algebra of alternating current theory: a + jb. Because four-terminal networks happen to have properties which are natural to matrices and because the recurring structure of matrix operations makes them well suited for performance on programmable calculators or small computers, it makes sense for the Amateur to apply matrix algebra to network design.

By Forrest Gehrke, K2BT, 75 Crestview, Mountain Lakes, New Jersey 07046

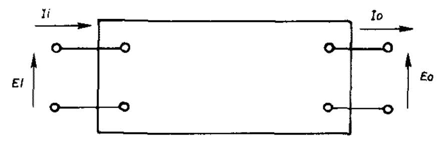


fig. 1. Four-terminal network showing reference directions for voltage and current. Subscripts i and o imply 'in' and 'out', respectively.

As with many other mathematical concepts, the application of matrix methods does not necessarily require a complete understanding of the underlying theory. It is in this context — i.e., in explanation of the *use* of matrices — that I address this subject. (For those who may wish to explore matrix algebra in depth, a brief bibliography is supplied at the end of this article.)[2]

Some fundamentals, such as the assumptions and restrictions and the notation employed, must be understood. Before all else, it should be emphasized that we are dealing only with alternating current *steady* state. (Matrix methods can be applied to transient and pulsed states, but that is outside the realm of this discussion.)

Four-terminal networks are a special form of a general network having a pair of input terminals and a pair of output terminals. Pictured in **fig. 1** is a box with the two pairs of terminals showing the reference directions for voltage and current. We may not know what is in the box, but we will suppose it to consist of any number of circuit branches and impedances with the following restrictions applying:

1. All impedances may be complex, but are linear and constant (time-invariant).

2. The network is passive, i.e., the only generators must be external sources (no dependent or internal sources), represented by E_i or E_o operating alone, or both simultaneously.

Despite not knowing the exact circuit inside the box, enough information can be deduced from measurements (amplitude *and* phase) made at the four terminals to produce a simple Tee or π circuit which is equivalent to it at any one frequency. We do this by defining a set of matrix parameters as follows:

$$A = (E_i/E_o) \qquad \text{with } I_o = 0 \tag{1}$$

Voltage E_i is applied to the input terminals and voltage E_o is measured at the output terminals with no load connected to the output.

$$B = (E_i/I_o) \qquad \text{with } E_o = 0 \tag{2}$$

Voltage E_i is applied to the input terminals and the short circuited output current I_o is measured.

$$C = (I_i/E_o) \qquad \text{with } I_o = 0 \tag{3}$$

Applying a voltage to the input terminals, measure input current I_i and output voltage E_o with no output load connected.

$$D = (I_i/I_o) \qquad \text{with } E_o = 0 \tag{4}$$

Applying a voltage to the input terminals, measure input current I_i and measure the short-circuited output current I_o.

The coefficients A, B, C, D are called general network parameters. Two relationships exist between inputs and outputs of a four-terminal network involving these parameters:

$$E_i = AE_o + BI_o \tag{5a}$$

$$I_i = CE_o + DI_o \tag{5b}$$

A and D are dimensionless transfer ratios, but B and C have the dimensions of impedance and admittance, respectively. In addition, a specific relationship exists between the network parameters because of reciprocity:

$$AD - BC = 1 \tag{6}$$

If we know any three of the four parameters, we can calculate the fourth. On the other hand, if we believe we know all four, this relationship gives us a means of verification, for if calculations using **eq. 6** do not hold, there must be an error.

The matrix used to describe four terminal networks is known as a square or network matrix. The network matrix is always portrayed this way:

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} \tag{7}$$

If the network contains no resistances, i.e., is lossless, A and D are real numbers and B and C are pure imaginary numbers (B and C carry the 'j' operator).*

Several kinds of matrix parameters such as S, H, and ABCD, have been developed for solving particular problems. Though one system is sometimes preferred over the other, it is possible to convert any of these parameter types to any other type, and to use *any* type of matrix parameter in the calculations discussed herein.

**Complex algebra, real* and *imaginary*, are mathematician's terms. For people working with electronics, these terms may unfortunately convey meanings which are not intended literally. The algebra, although different, *is not complicated*; the reactances resulting from inductances and capacitances *are neither unreal nor imaginary in effects*. However, the terms have been with us a long time and are here to stay.

Matrices may be manipulated — added, multiplied, inverted, reversed, partitioned — in many ways, but only according to special rules of procedure and order. For example, when four-terminal networks are cascaded, the effect may be calculated using matrix multiplication. If the individual matrix of each component network is known, the product is a new matrix of ABCD parameters for the overall network chain, allowing calculation of input/output relationships directly, end-to-end. If there is no need to determine the intermediate voltages and currents of the component networks, this is the way to go. (The order of this matrix multiplication is important. Obviously, there will be a great difference in results if the position of a 75-ohm 1/4 wave transformer followed by any length of 50-ohm line, is reversed!)

A four-terminal network is reversible if the A parameter is equal to the D parameter at all frequencies. This is another special case of the network matrix; we may reverse the connections to such a network without causing any change. Physically, a length of coaxial transmission line, a symmetrical Tee or π circuit are examples of such networks. The matrix of reversible networks appears like this:

$$\begin{bmatrix} A & B \\ C & A \end{bmatrix}$$

$$\text{with } A = \pm\sqrt{1 + BC}$$

When designing four-terminal networks, it is often necessary to express the current and voltage at the input side as a linear transformation of the current and voltage on the output side. For instance, with phased arrays we usually know what is wanted on the output side. What we need to know is the current and voltage required on the input side to produce those specific output conditions. Rearranging **eq. 5**, we can state several useful relationships. If each equation is divided by I_o we have:

$$E_i/I_o = (AE_o/I_o) + B$$

and

$$I_i/I_o = (CE_o/I_o) + D$$

E_o/I_o describes an external load which I will call Z_a. (I have chosen to use the subscript 'a' to avoid confusion with the commonly used 'o' which refers to a transmission line characteristic impedance.)

If the first of the above equations is divided by the second, and Z_a substituted for E_o/I_o, we obtain:

$$E_i/I_i = Z_{in} = (AZ_a + B)/(CZ_a + D) \qquad \textbf{(8)}$$

which defines the input impedance of the network in terms of the output load Z_a and the parameters of a network.

This leads to additional useful relationships:

$$E_o/E_i = Z_a/(AZ_a + B) \qquad \textbf{(9)}$$

$$I_o/E_i = 1/(AZ_a + B) \qquad \textbf{(10)}$$

$$I_o/I_i = 1/(CZ_a + D) \qquad \textbf{(11)}$$

$$I_i = I_o(CZ_a + D) \qquad \textbf{(12)}$$

$$E_i = I_o(AZ_a + B) \qquad \textbf{(13)}$$

If we know the input impedance and want to calculate the load impedance:

$$Z_a = (DZ_{in} - B)/(A - CZ_{in}) \qquad \textbf{(14)}$$

If the matrices of the fundamental types of four-terminal networks are known, along with the input or output impedance, all other network characteristics can be computed. Notice the recurrence of the terms $AZ_a + B$ and $CZ_a + D$ in the above relationships. If we program only these two calculations, we have substantially reduced the tedium of network calculations. (Most scientific calculators include the functions rectangular-to-polar and polar-to-rectangular, which takes care of most of the rest of the computations.*) Notice also that the ABCD parameters define the network operating characteristics, independent of the type of network. Therefore, if a length of coax and a π circuit or a Tee circuit have identical parameter values, the circuits will be exactly equivalent (even though the equations for calculating the parameters are different for each of these network types). Though not *all* types of networks may be transformed from one form to another, it is true often enough not to ignore the possibility. If true, use it to simplify your design.

Part 4 showed the basic building block four-terminal networks most useful to design of the drive networks for phased arrays. Presented in **table 1** are the parameter equations for each of those circuits, using the same notation. Note that this discussion is confined to the lossless case, since for low band frequencies losses are usually negligible. In the more general case, the procedure is still valid with loss terms introduced. However, it's not needed in this discussion.

using the ABCD parameters

Example calculations, which usually improve understanding, also help illustrate the versatility of matrix methods. I will first show the relatively simple case of a quarter-wave transformer and then proceed to design a real network for a 2-element array.

*For those familiar with Hewlett-Packard calculators and RPN, an SASE to the author will bring a 98-step program developed for the HP-19C which can calculate eqs. 8 through 13 using complex algebra. Translation for other programmable H-P calculators should not be difficult.

The electrical length of a quarter-wave transformer is 90 degrees. If we are not interested in circuit components values, we do not need to know the frequency. What is required is the angular length and the characteristic impedance of the circuit. "Electrical length" (in any units of length) is always defined as the length under *matched* load conditions. But this does not imply that the current or voltage phase displacement at other than matched conditions is necessarily equal to the electrical length of the circuit.[3] The quarter-wave transformer is an exception; this consequently accounts for its great utility. As long as the load is a *pure* resistance, the current and voltage phase displacement is 90 degrees even though not matched. If our transformer is made from a 50-ohm transmission line, it has the following ABCD parameters:

$$A = \cos 90^\circ + j0 = 0 + j0$$

$$B = 0 + j50 \sin 90^\circ = 0 + j50$$

$$C = 0 + \left(\frac{j \sin 90^\circ}{50}\right) = 0 + j0.02$$

$$D = \cos 90^\circ + j0 = 0 + j0$$

Assume that the load is a pure resistance of $35 + j0$:

$$AZ_a + B = 0 + j50$$

$$CZ_a + D = 0 + j0.7$$

and $Z_{in} = (AZ_a + B)/(CZ_a + D)$
$= 71.4285 + j0 \text{ ohms}$

Though we already know the current phase displacement, we can also determine the current amplitude ratio, a factor often required in antenna array calculations:

$$I_o/I_I = 1/(CZ_a + D) = 1.4285 \angle -90^\circ$$

Assuming the load current, I_o, to be $1 + j0$, the input voltage is $E_i = I_o\ (AZ_a + B) = 50\ \angle 90^\circ$, another value often needed in array network design.

In short, with **eqs. 8** through **13**, (in some cases assuming values for output current or voltage), we can find any of the input conditions using the ABCD parameters of the circuit.

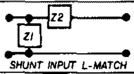
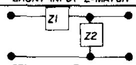
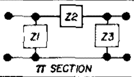
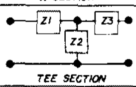
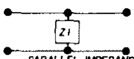
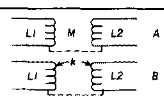
Suppose we had chosen a π network for our quarter-wave transformer. If we use a reversible π circuit designed to be coax-equivalent we know that[1]

$$X_1 = X_3 = -(Z_o \sin \theta)/(1 - \cos \theta)$$

and $X_2 = Z_o \sin \theta$

where Z_o and θ is defined the same way as it is for coax. As already indicated, the ABCD parameters define the circuit characteristics. Since the circuit characteristics are supposed to be the same, we

Table 1. Four-terminal network block diagrams and associated matrix forms.

No.	NETWORK	MATRIX $= \begin{bmatrix} A & B \\ C & D \end{bmatrix}$
1	 SHUNT INPUT L-MATCH	$\begin{bmatrix} 1 & Z2 \\ 1/Z1 & (Z1-Z2)/Z1 \end{bmatrix}$
2	 SERIES INPUT L-MATCH	$\begin{bmatrix} (Z1+Z2)/Z2 & Z1 \\ 1/Z2 & 1 \end{bmatrix}$
3	 π SECTION	$\begin{bmatrix} \frac{(Z2+Z3)}{Z3} & Z2 \\ \frac{(Z1+Z2+Z3)}{Z1\ Z3} & \frac{(Z1+Z2)}{Z1} \end{bmatrix}$
4	 TEE SECTION	$\begin{bmatrix} \frac{(Z1+Z2)}{Z2} & \frac{(Z1Z2+Z1Z3+Z2Z3)}{Z2} \\ \frac{1}{Z2} & \frac{(Z2+Z3)}{Z2} \end{bmatrix}$
5	 SERIES IMPEDANCE	$\begin{bmatrix} 1 & Z1 \\ 0 & 1 \end{bmatrix}$
6	 PARALLEL IMPEDANCE	$\begin{bmatrix} 1 & 0 \\ \frac{1}{Z1} & 1 \end{bmatrix}$
7	 TRANSMISSION LINE	$\begin{bmatrix} \cos\theta & jZo \sin\theta \\ j\frac{\sin\theta}{Zo} & \cos\theta \end{bmatrix}$
8A 8B	 COUPLED CIRCUITS	(A) $\begin{bmatrix} \frac{L1}{M} & \frac{j\omega(L1L2-M^2)}{M} \\ \frac{-1}{\omega M} & \frac{L2}{M} \end{bmatrix}$ (B) $\begin{bmatrix} \frac{1}{k}\sqrt{\frac{L1}{L2}} & \frac{j\omega\sqrt{L1\,L2}(1-k^2)}{k} \\ \frac{1}{j\omega k\sqrt{L1\,L2}} & \frac{1}{k}\sqrt{\frac{L2}{L1}} \end{bmatrix}$

should expect to find the parameter values to be identical even though calculated differently.

$$A = \frac{0 + j\,50 \sin 90^\circ - j\,\frac{50 \sin 90^\circ}{1 - \cos 90^\circ}}{0 - j\,\frac{50 \sin 90^\circ}{1 - \cos 90^\circ}} = 0 + j0$$

$$B = 0 + j50 \sin 90^\circ = 0 + j50$$

$$C = \frac{0 - j\,\frac{50 \sin 90^\circ}{1 - \cos 90^\circ} + j\,50 \sin 90^\circ - j\,\frac{50 \sin 90^\circ}{1 - \cos 90^\circ}}{\left(0 - j\,\frac{50 \sin 90^\circ}{1 - \cos 90^\circ}\right)\left(0 - j\,\frac{50 \sin 90^\circ}{1 - \cos 90^\circ}\right)} = 0 + j0.02$$

$$D = \frac{0 + j\,50 \sin 90^\circ - j\,\frac{50 \sin 90^\circ}{1 - \cos 90^\circ}}{0 - j\,\frac{50 \sin 90^\circ}{1 - \cos 90^\circ}} = 0 + j0$$

Since the ABCD (parameter) values are identical, all other circuit relationships will be identical also. A similar parameter computation can be carried out with the coax-equivalent Tee circuit where[1]

$$X_2 = -Z_o/\sin\theta \text{ and } X_1 = X_3 = Z_o\frac{(1 - \cos\theta)}{\sin\theta}$$

Again the ABCD parameter values will be identical to the first two cases, though their computation requires yet another set of equations from **table 1.**

For the specific quarter-wavelength example we did not have to perform all these calculations to find the input impedance. We know all we need to know from the quarter-wavelength relationship,

$$Z_o = \sqrt{Z_{in}Z_a}$$

but when examining a new procedure it is always reassuring to be able to verify it with a more familiar one. Even for this example, we would not be able to calculate voltage or current input conditions quite so easily if the load were reactive. Best of all, matrix methods are applicable to any circuit without restrictions placed on electrical length. To illustrate this point, let's design a no-compromise feed network for a 2-element vertical phased array. Assuming directional switchability is desired, physical symmetry will dictate equal length element feeders. However, these need only be long enough to meet at a central switching point in the array. At a design frequency of 3.8 MHz, *the array in this example has* the following characteristics:

Equal amplitude current drive with a 90° phase displacement between elements. The elements are quarter-wave spaced and quarter-wave resonant.

From part 3 of this series[4] the driving-point impedances are:

$$Z_1 = 21.4 - j15 \text{ for element 1}$$

$$Z_2 = 51.4 + j15 \text{ for element 3}$$

Note: These impedances are for elements with an extensive ground plane.

At 3.8 MHz the element spacing is 64.71 feet. Allowing for some variation in placement of the switching relays and feed network, each feeder is arbitrarily cut to 34 feet. Assuming a line characteristic impedance, Z_0, of 50 ohms and velocity factor of 0.66 the electrical length of the feeders is 71.65°. The drive network will be matched to a 50-ohm line.

For a matched array we *must first determine* the resistance loads each network chain presents to the shack line. Assuming 1 ampere flowing into each element, the total power going to the array is the sum of I^2R inputs, or

$$21.4 \bullet 1^2 + (51.4) \bullet 1^2 = 72.8 \text{ watts}$$

table 2. Input conditions and ABCD parameters at each circuit junction with 1 ampere flowing into each element.

element 1		**element 2**	
Z_1	= 21.4 − j15	Z_2	= 51.4 + j15
E_1	= 26.134 $\angle -35.028°$	E_2	= 53.544 $\angle -73.731°$
I_1	= 1 $\angle 0°$	I_2	= 1 $\angle -90°$
71.65° coax		**71.65° coax**	
A	= 0.3148 + j0	A	= 0.3148 + j0
B	= 0 + j47.458	B	= 0 + j47.458
C	= 0 + j0.01898	C	= 0 + j0.01898
D	= 0.3148 + j0	D	= 0.3148 + j0
$AZ_a + B$	= 6.7372 + j42.7351	$AZ_a + B$	= 16.1819 + j52.1798
$CZ_a + D$	= 0.5995 + j0.4062	$CZ_a + D$	= 0.0300 + j0.9757
Z_1	= 40.7999 + j43.6326	Z_2	= 54.9379 − j14.9217
E_1	= 43.2629 $\angle 81.0410°$	E_2	= 54.6314 $\angle -17.2296°$
I_1	= 0.7242 $\angle 34.1195°$	I_2	= 0.9761 $\angle -1.7656°$
shunt L-match (leading)		**shunt L-match (lagging)**	
A	= 1 + j0	A	= 1 + j0
B	= 0 − j116.2630	B	= 0 + j45.0951
C	= 0 − j0.01047	C	= 0 + j0.0079
D	= −0.21678 + j0	D	= 0.64378 + j0
$AZ_a + B$	= 40.7999 − j72.6303	$AZ_a + B$	= 53.9379 + j30.1733
$CZ_a + D$	= 0.2398 − j0.4270	$CZ_a + D$	= 0.7616 + j0.4260
Z_1	= 170.0934 + j0	Z_2	= 70.8171 + j0
E_1	= 60.3324 $\angle -26.5554°$	E_2	= 60.3324 $\angle 27.4573°$
I_1	= 0.3547 $\angle -26.5554°$	I_2	= 0.8519 $\angle 27.4573°$
pi circuit (lag 54.0126°)			
A	= 0.5876 + j0		
B	= 0 + j137.6306		
C	= 0 + j0.004757		
D	= 0.5876 + j0		
$AZ_a + B$	= 99.9479 + j137.6303		
$CZ_a + D$	= 0.5876 + j0.8091		
Z_1	= 170.0934 + j0		
E_1	= 60.3324 $\angle 27.4572°$		
I_1	= 0.3547 $\angle 27.4572°$		

The voltage at the common connection of the array for matched conditions, i.e., a 50 ohm load, is $E = \sqrt{RW} = \sqrt{(50)(72.8)} = 60.3324$ *volts*. Since the element networks will be transformed individually to pure resistances whose paralleled value is 50 ohms, we must find the individual values. Going back to the resistive components of each driving-point impedance (and knowing the array's impressed voltage amplitude when correctly driven), we can determine what these resistive loads must be. Using the relationship $R = E^2/W$, these transformed loads are respectively:

$$(60.3324)^2/21.4 = 170.0935 \text{ ohms, element 1*}$$

$$(60.3324)^2/51.4 = 70.8171 \text{ ohms, element 2}$$

*Check whether these resistances when paralleled equal 50 ohms.

As recommended in part 4, the simplest network often results if the drive network for the $-90°$ phased element establishes the voltage amplitude and phase for the array common connection. Proceeding on that basis we will design the network for element 2 first. We need to transform the driving-point impedance of this element to the input end of its coax feeder and to determine the input voltage and current that must exist there. **Table 2** lists the input conditions for each element at each junction point and the ABCD parameters for each circuit. For simplicity, 1 ampere is assumed to be flowing into each element.

The design procedure for the lumped-constant part of the network transforms element 2 driving-point impedance, as seen at the input to its coax feeder, to the resistive load required, using a lagging phase shunt L-match. (The design equations for all discussed circuits are found in part 4.) This fixes the voltage amplitude and phase required for the array. Element 1 driving-point impedance, as seen at the input to its coax feeder, is transformed, with a leading phase shunt L-match, to the resistive load required for this chain. At the input to this L-match, the voltage amplitude is correct but the phase displacement has overshot the objective. (L-match circuits can be designed to produce either a specific resistive input or a specific phase displacement — not both.) The solution is to add a lagging π coax-equivalent circuit with a characteristic impedance of 170.0935 ohms and an electrical length equal to the difference between the phase angle existing at the input to this L-match ($-26.5554°$) and the angle required at the common connection point ($27.4572°$). This difference, i.e., the total angular displacement between these two vectors, is $54.0126°$. This phase correction circuit can be thought of as though we had somehow magically obtained approximately 26 feet of coax delay line having a characteristic impedance of 170 ohms. When doing chained network calculations, don't forget that Z_a (output load) for the circuit being computed is the *input* impedance of the preceding junction (looking towards the load). For example, when transforming the element driving-point impedance to the input end of its feeder, then the output load, Z_a, is the element's driving-point impedance. But when computing input/output relations for the succeeding L-match circuit, Z_a is now the impedance seen at the input end of the transmission line feeder, and so on.

final 2-element network design

The resulting design for the feed network of this array requires three inductances and four capacitances as seen in **fig. 2**. The component values are quite realizable. Evaluation of network designs is admittedly somewhat subjective. It is conditioned by the number of individual network circuits required, circuit component values (e.g., at 3.8 MHz lossless air core series arm inductances greater than 10 μH become physically large; shunt arm capacitance values less than 50 pF require more rigorous assessment of the unavoidable stray circuit capacitances). Should awkward values of components result from a particular design, it is often possible that using a different element to establish the voltage amplitude and phase at the array common connection point results in more physically realizable components.

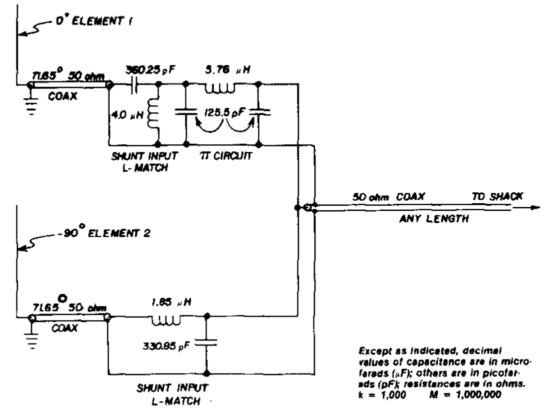


fig. 2. Feed network schematic for quarter-wave spaced two-element vertical array.

In a concluding article of this series I will discuss practical array and feed network construction and measurements.

This information, gleaned during the various phases of the development of my vertical phased arrays, should help the reader convert the theory presented on these pages into an actual physical structure — an antenna array that works the way it was designed to work.

references

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2. Frank Ayres, *Matrices*, (Shaum's Outline Series), McGraw-Hill Book Company; Noble and Daniel, *Applied Linear Algebra*, Prentice-Hall Inc.; L.A. Pipes, *Applied Mathematics for Engineers and Physicists*, McGraw-Hill Co.
3. Forrest Gehrke, K2BT, "Vertical Phased Arrays: part 1," *ham radio*, May, 1983, page 18.
4. Forrest Gehrke, K2BT, "Vertical Phased Arrays: part 3," *ham radio*, July, 1983, page 26.

ham radio

lightning and electrical transient protection

A variety of devices safeguard delicate electronic circuitry

It's a summer afternoon; the air feels heavy, dense. Towering gray-black cumulonimbus clouds advance across the sky. Suddenly lightning knifes from cloud to ground, filling the air with thunder. Bolts leap from cloud to cloud, cloud to ground, and from top to bottom within individual clouds. Rain pours down and then, almost as abruptly as it began, the thunderstorm disappears.

Snug at home, a ham fires up his rig to keep a schedule on 20 meters. He's running a few minutes late, but that's better than operating in the middle of an electrical storm. To his dismay, he discovers that his radio can't transceive any more. It sits dead, emitting a quiet hiss on every band.

Unknown to our ham, his rig gave up the ghost five minutes before he turned on the power. The FET in the receiver front-end had its gate destroyed by a lightning-induced static discharge.

Every year, hundreds of hams have the same unfortunate, unnecessary experience. *All* radio equipment can be protected, easily and inexpensively, but before we can give our rigs complete protection, we must first understand the causes and nature of static discharges.

Thunderstorms develop when warm, moist air is forced to rise. This may be triggered by the meeting of warm and cold air masses, which is especially common in the Midwest and along the Gulf Coast and Eastern seaboard. Or it may be induced by terrain, as frequently happens in the mountainous West. Thunderstorms are most likely to occur in those areas favored by frontal development or terrain or both, as shown in **fig. 1**.

For whatever reason, rising air begins cloud formation at *lifting condensation level* where its temperature drops below the dew point. The larger the temperature differential between the surrounding atmosphere and the rising column of air, the greater the speed and violence of development.

An average cumulonimbus cell will reach upward 6 miles (9.7 km) and develop vertical air drafts with velocities of 15 feet per second. Violent cells may rise to 60,000 feet and create intense drafts with speeds over 200 feet per second.

After the formation of ice crystals, a charge separation occurs within the cloud. The top of the cloud becomes positive; the underside, negative. The reasons for this separation are varied: frictional contact, melting, freezing, inductive charge transfer, water drop breakup, ion attachment. These mechanisms may operate independently or together and turn the cloud into a giant static electricity generator.

The development of extreme potential differences within the cloud leads to atmospheric breakdown and electrical conduction. Lightning strokes begin within the cloud and often between clouds. While nine out of ten discharges during thunderstorms are of this type, it's that remaining ten percent that are of primary concern to us.

lightning development

Development of a ground stroke follows a definite sequence: first, a faintly luminous tendril descends from the cloud base. This is the *stepped leader*. A stepped leader moves downward in jerky spurts of 150 feet (45.7 meters) at speeds of 75 miles (120 km) per second. It may branch out, with a new fork occurring at one of its hesitation points. When this leader is within 100 feet (30.5 meters) of the ground, the local electrical field becomes strong enough to initiate discharge from the ground. This *ground streamer* moves upward in a forked pattern. With the connection of the stepped leader and ground streamer, a heavy current (typically 20,000 amperes)

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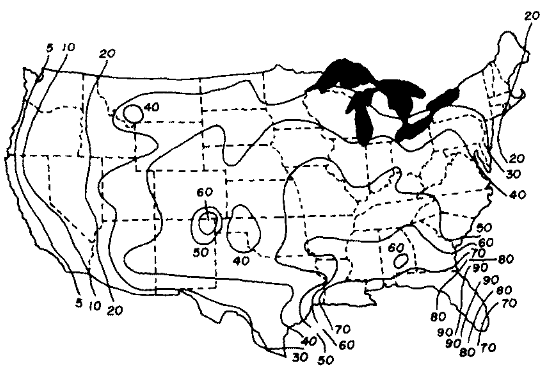

fig. 1. The mean number of thunderstorm days in the United States is related to both mountainous terrain and seasonal weather patterns. (Adapted with permission from *Understanding Lightning* by Martin Uman, copyright 1971, BEK Technical Publications, 1700 Painters Run Road, Pittsburgh, Pennsylvania 15243.)

flows from the ground to the cloud at a speed of 60,000 miles (97,000 km) per second. This *return stroke* produces the lightning flash that we observe.

Approximately 50 milliseconds after the return stroke, another tendril of luminosity descends from the cloud base. This is the *dart leader*, and it traverses the same channel followed by the return stroke. Its ground contact triggers a second return stroke of slightly less power than the first. Most cloud-to-ground discharges consist of three or four return strokes, each successively penetrating farther into the cloud to tap new charge accumulations.

The quantity of energy in an average lightning stroke can approach 1,000,000,000 joules. It transfers some 10 coulombs of charge across a potential difference of 100 million volts. This release of energy in less than 300 *microseconds* is equivalent to many megawatts of power. While lightning strokes are normally measured in thousands of feet, horizontal strokes up to 100 miles (161 km) in length have been observed. Similarly, voltages also vary, and potential differences of as high as one billion volts have been recorded. The sudden energy release of these lightning strokes can generate intense electrostatic and electromagnetic fields miles away from their origin.

lightning protection

Several steps are involved in protecting electronic equipment from lightning-induced static discharge. The first rule is to bond all metal objects to a ground system in order to provide a low resistance, high capacity reservoir for soaking up static charges. Pay particular attention to your tower since it will act just like a giant lightning rod: clean and lubricate all joints with conductive grease, then tie the tower to your ground system with heavy copper or aluminum wire. *Simply having the tower base in the earth is no guarantee of a good electical ground.* (Commercial radio towers use wide copper straps run in several directions for some distance to secure a good ground.) Properly grounded, a tower located close to the house can provide a degree of protection for the dwelling and occupants against lightning damage.

The often-repeated suggestion to disconnect antennas during a storm is still good advice. The problem, however, is that the operator may not be home to make this disconnection, or that the storm may be several miles away and out of sight. (see "10 GHz Weather Radar", *ham radio*, September, p. 61, for discussion of a method for detecting nearby thunderstorm activity.) Generally, by the time a ham thinks to unhook the coax, the damage is done.

other hazards

Several other sources of static electricity are common and may severely damage equipment. High winds, especially under conditions of very low humidity, can cause static in much the same manner as does dragging your feet over a carpet. Blowing powder snow may also cause static build-up on exposed metal surfaces.

Conventional lightning arrestors offer little protection for modern equipment because most of us operate *with solid-state rigs;* older tube-type radios were far more forgiving of static buildup than transistors or field-effect devices. Many tubes could actually arc over with minimal damage to either the tube or related components — but not modern transistors or, in particular, FETs. MOSFETs, particularly those without gate protection, can be irrevocably damaged by simply picking them up. Solid-state devices cannot tolerate any type of overvoltage on their base or gates, for even a microsecond, without sustaining damage. This damage may be total, in which case the receiver is dead; or it may be partial, leaving the operator thinking that propagation just isn't what it used to be.

Air gap lightning arrestors fail to protect equipment because the characteristics that determine their firing ability vary with both the air pressure and humidity of the environment. Under optimal conditons their response time is far too slow; even when these devices do fire, they retain a high enough voltage across the gap to damage sensitive equipment. Normally, 3 to 5 kilovolts is required to trigger the spark gap and when operating will still leave a 50-80 volt potential at the antenna terminal.

What we need is a device that will fire within nanoseconds and clip off high voltage spikes effectively. Two such devices are shown in **fig. 2.** Unlike conventional air gaps, these lightning protectors util-

fig. 2. While designed for indoor use, both Alpha Delta Transi-Trap and Drake Surge Shunt may be installed outdoors if appropriate weatherproof shielding is provided.

ize a hermetically sealed tube filled with Krypton 85, a gas of known breakdown characteristics. Since these devices are sealed, they remain unaffected by changes in humidity or atmospheric pressure. The tube design is such that their firing properties remain constant under all conditions. The gas conducts very effectively and reduces the voltage to values safe for solid-state devices. The typical firing speed is 100 nanoseconds and conduction begins at 750 volts for transient pulses.

Unlike old-style lightning arrestors, both of these mechanisms shunt the static electricity directly to ground rather than to the receiver chassis. Bypassing the receiver in this manner prevents the chassis potential from being raised to several thousand volts above ground level.

powerline transients

Receivers and transceivers can also be ruined by the effects of voltage transients coming in through the back end via the AC mains. Just as lightning can induce currents in the antenna system, it can also induce voltages in nearby power lines. These transients, running up to thousands of volts, can damage power supplies and internal components.

A lightning return stroke generates an intense RF field over a range of frequencies from kilohertz to Gigahertz. Most of the energy is centered around a frequency of 5 kHz, which makes long power lines very effective antennas. Since a lightning strike can generate electric fields of 60-80 volts per meter at distances of a mile or more, voltage transients of kilovolt magnitude can easily show up on power lines.

A second type of transient is created by a short circuit in the power distribution system. This may happen any number of times during the year, depending on where you live and how well the local power company maintains the system. The clearing of a short circuit is effected by blowing a fuse or, more commonly, by tripping a circuit breaker. This abrupt termination of current induces a high voltage transient on the power lines from the collapsing magnetic field of the power distribution transformers. Spikes of up to 6000 volts for periods of 5 to 10 microseconds are not uncommon, as shown in **fig. 3**.

Another source of voltage transients is inductive load switching of motors associated with furnaces, washing machines, and other appliances. When the current flow through motor windings is terminated, the magnetic field collapses and induces a reverse voltage of such magnitude that it may jump across the switch contacts to the power mains. Once an arc is established across the switch contacts, current flow continues to oscillate back and forth until there is insufficient energy to maintain the arc.

The problem of protecting your equipment against these transient voltages is further complicated by mode of transmission to the equipment. These

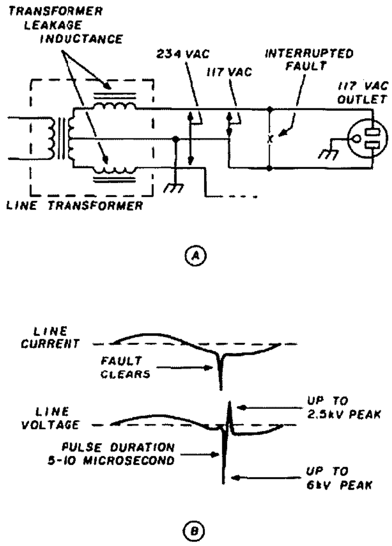

fig. 3. High voltage transients can occur on household 117 volt lines when a short circuit is cleared. (3A) shows circuit condition conducive to transient generation; (3B), current and voltage waveforms of spike. (Adapted from *QST*, February, 1982.)

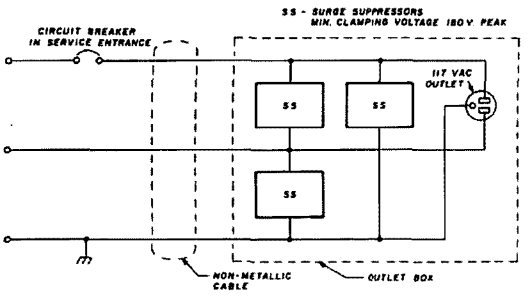

fig. 4. Three metal oxide varistors (MOV's) in a wall outlet provide maximum transient protection against the effects of both differential and common mode coupling within the house wiring.

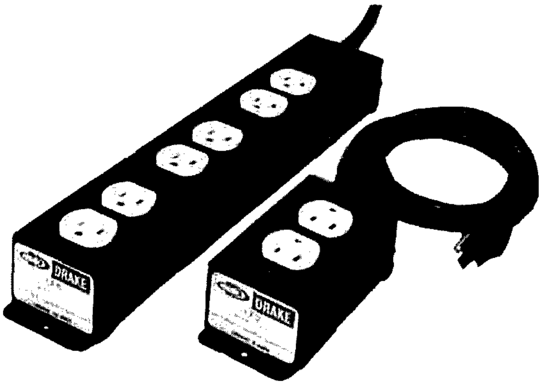

fig. 5. Specially-designed terminal strips available from Drake and other manufacturers offer protection from power line spikes and transients.

pulses can exist between individual lines or between lines and ground in the house wiring. A further complication is the inherent capacitance of the house wiring which allows these transients to move from one wire to another.

transient protection devices

Several devices that will protect equipment from these damaging voltage transients are available. The General Electric Home Lightning Protector, for example, is built to protect electrical devices within the home from lightning-induced voltage transients. Designed for mounting in either the service entrance box or the weatherhead for the incoming AC mains, it intercepts the transients before they enter the house wiring.

The metal-oxide varistor (MOV), also manufactured by General Electric, exhibits a resistance that is inversely proportional to the applied voltage. It is a non-polarized device for installation between wires or from wires to ground as shown in fig. 4. MOVs are built to respond to transients in nanoseconds and have clamping voltages ranging from 10 to 1500 volts. Most have continuous power ratings of 1/2 to 5 watts with peak dissipation of over 600 joules of energy. These may be installed in service entrances or within the equipment itself.

General Semiconductor Industries manufactures the TransZorb, a semiconductor device with a large PN junction for transient protection. TransZorbs are capable of handling very high power levels for short periods of time (100,000 watts for 100 nanoseconds).

Both TransZorbs and MOVs look like oversized disc ceramic capacitors and may be installed with the electronic equipment or more easily at the back of an AC receptacle (after switching off the circuit breaker at the service panel).

Alpha Delta, Drake, and others sell terminal strips with built-in protective devices (fig. 5). These may be purchased in various configurations depending upon the number of outlets desired. Radio Shack markets the Voltage Spike Protector for single-plug protection. It resembles an oversize AC plug and contains a GE MOV wired across the AC line. It can clamp a 435 volt, 50 ampere, 20 micro-second surge and is the least expensive of the commercially made units.

A list of manufacturers and distributors of lightning protection devices is provided in table 1. Unfortunately, none of these devices will protect your valuable equipment against a direct lightning strike. The

table 1. List of manufacturers of distributors of surge and transient protection devices.

Alpha Delta Communications P.O. Box 571 Centerville, Ohio 45459	Telex/Hy-Gain 9600 Aldrich Avenue South Minneapolis, Minnesota 55420
R.L. Drake Company 540 Richard Street Miamisburg, Ohio 45342	Kalglo Electronics 6584 Ruch Road East Bethlehem, Pennsylvania 18017
Electro Protection Devices, Inc. P.O. Box 673 Waltham, Massachusetts 02254	MFJ Enterprises, Inc. 921 Louisville Road Starkville, Mississippi 39759
Electronic Specialists, Inc. 171 South Main Street Natick, Massachusetts 10760	Newark Electronics 500 North Pulaski Road Chicago, Illinois 60624
Encomm, Inc. 2000 Avenue G, Suite 800 Plano, Texas 75074	Poly Phaser P.O. Box 2001 Kissimmee, Florida 32741
General Electric Company Semiconductor Division West Genesee Street Auburn, New York 13021	Technico, Inc. 9051 Red Branch Road Columbia, Maryland 21045
General Semiconductor Industries P.O. Box 3078 Tempe, Arizona 85281	Trio-Kenwood Communications 1111 West Walnut Compton, California 90220

old rule is still the best: during an overhead electrical storm, disconnect all antennas and unplug all equipment. And don't forget that the potential for static damage exists at any time; take appropriate protective measures. Much static damage is not immediately apparent, and it would be a shame to discover that your poor reception is not due to the decline in the sunspot cycle.

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PS Form 3526 June 1982

ham radio TECHNIQUES

Bill Orr W6SAI

intermodulation distortion

Ham radio equipment has come a long way in the last ten years. All you have to do is to pick up a ten-year-old copy of *ham radio* and compare the articles and advertisements of yesterday with those of today.

One of the interesting developments that has taken place during this decade is the gradual improvement in linearity and the reduction of intermodulation distortion. These improvements are particularly noticeable in some of the new transceivers and exciters available today. But before I discuss the improvements, it would be a good idea to review the fundamentals of intermodulation distortion (IMD) and its importance in HF communications.

Intermodulation distortion is a particularly unpleasant form of nonlinearity that should be of interest to those operators using linear amplifiers (either in their exciters or as an auxiliary unit). IMD occurs in a nonlinear device driven by a complex signal having more than one frequency.[1,2]

As speech is made up of a multitude of frequencies, and the perfect linear amplifier has yet to be built, IMD is of concern to every Amateur interested in voice or multiplex transmission.

A linear amplifier is an amplifier in which the output signal is an *exact* replica of the input signal. If the output signal does not duplicate the input signal, intermodulation distortion is created. IMD shows up on the air as a "roughening" of the signal accompanied by broad sidebands ("splatter"). Some Amateurs who are acutely sensitive to IMD can actually hear it as a growl on the low-frequency portions of voice transmissions and, in years past, could accurately identify the manufacturer of an SSB transmitter by merely listening to the signal on the air!

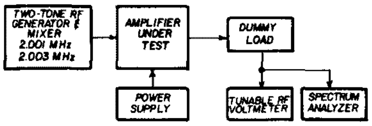

fig. 1. Intermodulation distortion analyzer. A two-tone RF generator with excellent IMD characteristics (third-order distortion products better than 70 dB down from one tone of a two-tone signal) drives the amplifier under test. A portion of the output signal is observed on a spectrum analyzer and measured by a tunable RF voltmeter. The test may take place at any frequency within the operating range of the amplifier. 2 MHz and 30 MHz are common test frequencies.

Intermodulation distortion may be examined and the amplifier tested by use of a distortion analyzer (**fig. 1**). An input signal consisting of two sine waves of equal amplitude (for example, 2.001 MHz and 2.003 MHz) is applied to the amplifier. The distortion products in a nonlinear stage appear as spurious signals within the passband of the amplifier and also at the harmonic frequencies, and the intermodulation distortion, as additional signals! In the example in **fig. 2**, the IMD products are located 2 kHz apart and occupy the span of 1995 kHz to 2009 kHz, or more. Thus, a signal that should be only 2 kHz wide is now 14 kHz wide, thanks to intermodulation distortion (**fig. 2**). The situation can be worse than this as a greater degree of distortion produces additional spurious signals on each side of the two desired signals.

IMD is independent of the operating frequency. If a 20-meter linear amplifier were tested for IMD at 14.200 MHz using a two-tone test signal whose tones were 2 kHz apart, IMD products would form a spectrum of signals, 2 kHz apart centered around 14.200 MHz. Photos of this type of distortion have appeared from time to time in product reviews.

the IMD rating

Intermodulation distortion is expressed in relation to the output level of an amplifier. The standard method of specifying the magnitude of the distortion products is to specify the reduction in decibels of one distortion product from one tone of a two-equal-tone signal. For example, if an amplifier (or tube or transistor) under a particular set of operating conditions has third-order distortion products of −30 dB, this means that the products have an amplitude 30 dB below one of the two test tones. It is *not* correct to compare one distortion product to the sum of the two tones; that is to say, the PEP value of the signal. If this is done, the resulting distortion figure is 6 dB "better" than the correct example (−30 dB rather than −24 dB). Unfortunately many product reviews and amplifier specifications take advantage of this "one-upmanship" because the better figure looks nicer in the advertisement!

IMD then and now

Has any improvement been made in the intermodulation distortion level of today's exciters and amplifiers, as compared to those products in 1973? I think there's been worthwhile improvement as new devices and circuit techniques have been developed.

Many early exciters used the so-called television "sweep tubes" as linear amplifiers. The popular 6LQ6 tube is a good example. This pint-size "bottle" could squeeze out up to 125 watts PEP at a plate potential of only 800 volts. It was an inexpensive and effective amplifier tube that attained wide acceptance in yesterday's equipment. Unfortunately, the third-order IMD products of the 6LQ6 when run in this fashion were only 18 dB below one tone of the two-tone power level. And if the signal from this amplifier were run through a second amplifier stage to bring it up to the legal Amateur power-input level, the IMD products would be brought up by 10 dB or so in addition to the wanted signal. The result would be a broad, fuzzy-sounding signal having about 15 watts of splatter power in the third-order distortion products. It may not sound like much, but 15 watts of unwanted splatter on top of the signal you're trying to listen to can be intolerable.

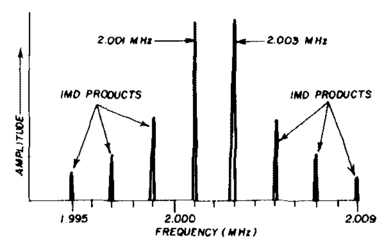

fig. 2. Two-tone IMD test signal has a 2 kHz spacing between the tones. In this representative case, the IMD products form an additional band of frequencies, spaced 2 kHz apart, centered on the test frequency. In this example, the third-, fifth- and seventh-order products are shown. The signal is 14 kHz wide. Higher-order products produced under severe distortion widen the signal even more than shown in this example.

The equally popular 6146 (6146B) was designed for RF services (as opposed to the 6LQ6), and it had much more appealing IMD ratings. At 800 volts and 60 watts power output, the IMD products were 24 dB below one tone of a two-tone test signal. That's 6 dB better than the 6LQ6, a worthwhile improvement.

The Collins Radio Company, now a division of Rockwell, pioneered the use of RF feedback to improve the linearity of an amplifier. It s use in the famous S-line transmitters provided an IMD level of better than −30 dB compared to one tone of a two-tone test signal.

And there the matter rested for a while. Perceptive Amateurs could pick out the S-line transmitter on the air, as it sounded immeasurably better than other equipment that did not incorporate RF feedback. However, as solid-state amplifiers came into general use, they were able to duplicate the IMD level set by the S-line as they, too, used fairly linear devices and an RF feedback loop. Thus, over the years, the "30-dB level" became an unofficial standard for measuring the excellence (or lack thereof) of IMD in Amateur equipment.

IMD today

Things are looking up. While IMD levels of −50 to −60 dB may be common in deluxe commercial and military gear, such levels are prohibitively expensive and too complicated for the competitive Amateur market. Despite this, improvements in intermodulation distortion in ham gear are here to stay. New solid-state devices skirt the −35 dB level with RF feedback, and at least one Amateur transceiver* has reached this favorable level using the common 6146B tube.

This is a far cry from the old sweep-tube "distortion generator." An IMD level of −35 dB sounds *clean* to the ear, even when the equipment is being run next door. Undoubtedly, significant improvements in IMD reduction will come about in the next few years, and I look forward confidently to an IMD level of −40 dB before the end of this decade.

Unfortunately, the good IMD rating of an exciter can be ruined quickly

*The Yaesu FT-102 HF transceiver boasts three 6146s in the ouput stage. Excellent linearity (better than −35 dB) is achieved by combining a relatively high resting plate current level with RF feedback.

by the careless operator who operates with "all knobs to the right." You hear plenty of these lids on the air today; will they be gone by 1990? I doubt it.

HF communication lives!

Remember the days when communications specialists predicted the demise of HF radio for long-distance communications? Why put up with static, fading, and interference on an unreliable HF circuit when you can do so much better with a reliable satellite link? In effect, HF radio was dead and buried by new communications techniques in the microwave satellite range. But times change! An article in the June, 1983, *Defense Systems Review* magazine[3] says it this way:

"Ten years ago it was thought that HF long distance communications for fleet operations would soon pass its heyday because of the capabilities of satellite communications. But the real vulnerability of satellite communications to antisatellite attack and potential propagation disruption with a nuclear detonation combined with the improved real-time propagation measurement capability give HF communications a new lease on life. The fact that HF communication is relatively inexpensive, flexible, and its assets are in place should give rise to increased interest to improve HF architecture and capabilities, relatively neglected for far too long."

Well, hurrah! The ionosphere has been rediscovered for long-distance communications as "a satellite that doesn't fall down." Spearheaded by the U.S. Navy (who started using HF communications in approximately 1911), communicators are starting to re-examine the HF spectrum as an alternative to satellite circuits.

CW lives, too!

Did you see the full-page in the September issue of *ham radio*, placed by the Central Intelligence Agency, offering openings for electronics technicians and communications and radio operators *with Morse Code ability preferred*? I wonder what the CIA thinks of FCC pressure to dilute Amateur Radio with no-code licensees?

a simple antenna for 21-28-50 MHz

Some new triband transceivers on the market cover only the 21-, 28-, and 50-MHz bands. A simple triband antenna for these ranges is shown in **fig. 3**. These dipoles are paralleled at the feedpoint and fed with a coaxial line. The dipoles are trimmed a bit to provide the lowest SWR figure at your chosen point in each band. A simple transmatch placed at the station end of the line may ease loading problems.

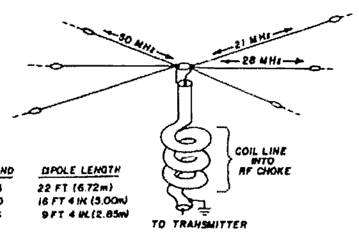

fig. 3. Triband dipole for 15, 10, and 6 meters. Separate the tips of the dipole at each end as far as possible. Transmission line is made into a coil below feedpoint to isolate outer surface of line from field of antenna. For RG-58/U line, use four turns, wrapped into a coil about six inches (30 cm) in diameter.

moonbounce revisited!

Another printing of the *Moonbounce Notes* is at hand. Send four first-class stamps (or four IRCs) to me at EIMAC, 301 Industrial Way, San Carlos, California 94070 and I'll send this interesting information to you.

references

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2. William I. Orr, W6SAI, "Intermodulation Distortion in Linear Amplifiers," *QST*, September, 1963. (For a reprint of this article, write author and request bulletin AS-21. Include two first-class stamps).
3. "High Frequency Transmissions Enjoy A Renewed Life," *Defense Systems Review*, June, 1983, published by Cosgriff-Martin & Cutter, Inc., Box 2828, Santa Clara, California 95055.

ham radio

the
weekender

In New England they say, "If you don't like the weather, wait a minute." NOAA lets you know what to expect

2-meter weather converter

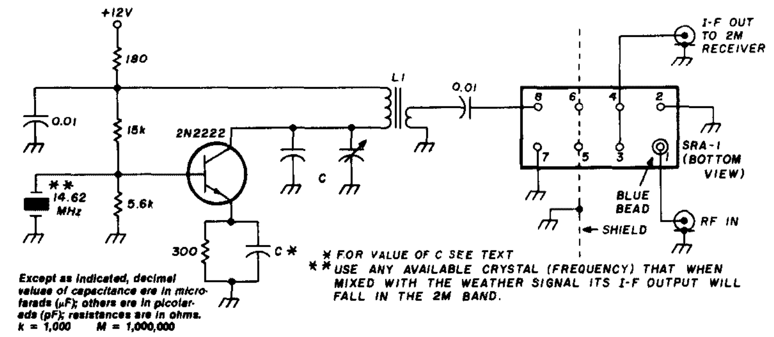


fig. 1. Weather converter consists of a 14.62 MHz local oscillator and a $10 double-balanced mixer.

Because I sometimes enjoy listening to the VHF public service bands and the NOAA weather broadcasts at 162 MHz, I designed and built a simple converter to convert these frequencies to 2-meter FM for reception on my transceiver.

The 162.40 MHz signal mixes with a 16 MHz local oscillator to produce a signal at 146.40 MHz. A typical converter has four stages: the RF amplifier, mixer, output filter, and local oscillator. Because a mixer is a multiplying device, it produces a signal at both the sum and difference of the input and the local oscillator frequencies. Conversely, for given intermediate and local oscillator frequencies, two input frequencies may be received. Tuned RF stages eliminate all but the desired frequencies.

I chose a passive double balanced mixer (DBM) for this converter. It is an integrated component with a frequency response from DC to 500 MHz. I used a Mini Circuits Lab SRA-1 ($9.95, in small quantities) that requires oscillator injection of about +7 dBm (5 mW) and has an inherent input impedance of 50 ohms. If IF feedthrough is about −45 dB and conversion loss is 6 dB, it is ideal for mixers when close frequencies are used.

By Ladimer S. Nagurney, WA3EEC, 73 Blackberry Lane, Amherst, Massachusetts 01002

The schematic diagram shown in **fig. 1** is the complete circuit. The local oscillator uses a 2N2222 that delivers about +10 dBm. The primary of L1 is selected to have a reactance of about 200 ohms at the crystal frequency. For example, at 15 MHz L1 is 20 turns on a T50-2 toroidal core. The secondary is 4 turns. The combination of the fixed and variable capacitors at the collector should resonate with L1. The emitter bypass capacitor should be about the same value but need not be adjustable. For 15 MHz, a 20 pF fixed and 7-45 pF variable capacitor were used in the collector and a 39 pF fixed value capacitor was used in the emitter. I used a 14.62 MHz crystal from my junkbox and tuned my transceiver to 147.78 MHz in order to receive 162.40 MHz.

The whole circuit was built on a small piece of PC board and mounted in a minibox. No pattern was etched on the board. The DBM was mounted by drilling a grid of eight holes at one end. A shield of PC material was used to separate the local oscillator from the IF and RF. The holes for the ungrounded pins of the DBM were drilled larger for clearance. A blue bead indicates pin 1 of the DBM.

The local oscillator was built using isolated pad

techniques. An approximately 3/4-square inch piece of PC board was divided into four equal pads with a hacksaw to cut the foil. This was placed upon the groundplane on the same side as the side on which the DBM pins protrude. It is secured by the resistor and capacitor leads soldered between the various pads and ground. The local oscillator was tested before the DBM was installed by attaching a 47-ohm resistor to the output capacitor as a load and using an oscilloscope to observe the signal. The output was clearly heard on my communications receiver.

The IF output of the mixer utilizes the excellent selectivity capabilities of the front end of a typical 2 meter Amateur transceiver. Since the input of the transceiver is matched to 50 ohms, no additional impedance matching was necessary.

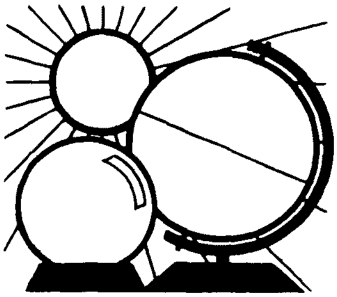
Note no RF filtering is used at the input. Because I could find no interfering signals near 133.16 MHz, I just connected the antenna directly to the input. If one wanted to eliminate this image, a highpass filter with a cutoff below 150 MHz could easily be added. Similarly, a lowpass filter could be used to eliminate the higher image if one wanted to copy a signal below the transceiver frequency.

Even though the mixer exhibits some loss, I found it to be acceptable for local listening. As soon as power was applied, the Providence NOAA weather station was received with full quieting occurring. Either a 162 MHz preamp before or a 2-meter preamp after the converter could be used to eliminate conversion losses.

If one were to choose a common repeater frequency for the IF, a receiver tuned to the weather channel could copy it because of the reciprocity of the mixer.

A note of caution: although DBMs are rugged, they will usually not withstand transmitter power levels. Also, the RF filtering in this simple converter is such that one might put out *several* signals if the unit were used for transmitting.

ham radio



DX FORECASTER

Garth Stonehocker, KØRYW

ray tracing

If Santa were to bring you a ray tracer DX machine for Christmas, you'd have the ultimate DX forecaster right in your home.

Ray tracing between transmitter and receiver locations *is* possible; see this column in the October issue for a diagram showing a trans-equatorial ray path (signal) traced through the ionosphere to give one-long-hop propagation.

Another example of ray tracing technique is to search the ionosphere between your transmitter and the DX QTH, varying the take-off angle, bearing and frequency. (A computer could quickly determine the best path to use and specify these three corresponding parameters.) Ray tracing procedure consists of the following steps:

1. Determine the electron (ion) density at heights above the earth along the ray (signal) path using ionosonde, rocket, or satellite measurements.

2. Generate ionospheric contours using the electron density versus height data.

3. Trace the ray (path) through these contours starting from the transmitted signal. The changing refractive indices along the path determine a new ray direction.

4. Follow the ray as it bends from one point to the next until it emerges from the ionosphere at an angle heading down to earth toward the receiver.

All these steps are depicted in the diagram.

Ray tracing is not commonly done because of the limited availability of electron density profiles around the world. Even if the profiles were available, a substantial computer would be needed for ionosonde data conversion and the electron density profiles storage. However, Radio Amateurs are now using more sophisticated computers, and ionospheric measurements can be made with SSB equipment; maybe someday you *will* be able to get a ray tracing machine for Christmas.

iast-minute forecast

December is one of the winter DX season months exhibiting low thunderstorm QRN and low probability of geomagnetic disturbances or at least long quiet periods in between disturbances. However, this year in the 11-year sunspot cycle may see a few more disturbances than in other years. Expect disturbances around December 4, 14, 17, 22, and 27. If WWV is broadcasting a radio flux greater than 140, good trans-equatorial openings might occur on 10-30 meters. If the flux is lower than 140, expect disturbances to last longer and fades to occur on the lower bands on higher latitude east-west paths.

An annular eclipse of the sun is to begin at 0941 UT in northeast North America, traveling across Iceland to England, Southern Europe, North Africa and ending in Southwest Asia at 1520 UT. The full moon is on the 20th and perigee on the 22nd. Winter solstice is on the 22nd at 1030 UT.

The Geminid meteor shower, which peaks on December 13 and 14, provides the richest and most reliable display of the year, with rates of 60 to 70 meteors per hour (determined mainly by radio, because of the poor weather in December). Also, a smaller portion of the shower (15 to 20 per hour) will be observed on December 22.

band-by-band summary

Ten, fifteen, twenty, and thirty meters will have DX openings from most areas of the world during daylight and into the evening almost every day. Long skip and one-long-hop trans-equatorial openings will occur in the early evening hours. These openings should occur during periods of disturbed geomagnetic field. Otherwise watch for high solar flux days for ten and fifteen meter openings.

Thirty, forty, eighty, and one-sixty meters are the night DXer's bands. Excellent extended periods of long skip, shorter than on the higher bands, will occur. Low noise and quiet geomagnetic conditions generally result in pleasant operating this time of year. Happy Holidays, and lots of DX during the coming New Year!

ham radio

DECEMBER

WESTERN USA

GMT	PST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	4:00	20	30	15	10	15	10	10	20
0100	5:00	20	30	15	10	15	10	10	20
0200	6:00	20	30	15	10	15	10	10	20
0300	7:00	20	30	15	10	15	10	10	20
0400	8:00	20	30	15	10	15	10	10	20
0500	9:00	*20*	*30*	20	10	15	10	15	20
0600	10:00	*20*	*30*	20	10	15	10	15	20
0700	11:00	*20*	*30*	20	15	15	15	15	30
0800	12:00	*30*	*30*	20	15	*20*	15	20	30
0900	1:00	*30*	*40*	20	*15*	*20*	15	20	*30*
1000	2:00	*30*	*40*	*20*	*20*	*20*	20	20	*30*
1100	3:00	*30*	*40*	*20*	*20*	*20*	20	20	*30*
1200	4:00	*30*	*40*	*20*	*20*	*20*	20	*20*	*30*
1300	5:00	*30*	*30*	20	*20*	*20*	*20*	*20*	*30*
1400	6:00	*40*	*30*	15	*20*	*20*	*20*	*20*	*30*
1500	7:00	*40*	*30*	15	*20*	*15*	*20*	*20*	*40*
1600	8:00	*40*	*30*	15	20	*15*	15	*20*	*40*
1700	9:00	40	20	15	20	*15*	15	15	*30*
1800	10:00	40	20	10	15	*15*	15	15	*30*
1900	11:00	40	20	10	15	15	10	15	20
2000	12:00	40	20	10	15	15	10	15	20
2100	1:00	30	20	10	15	15	10	10	20
2200	2:00	30	30	15	10	15	10	10	20
2300	3:00	30	30	15	10	15	10	10	20

MID USA

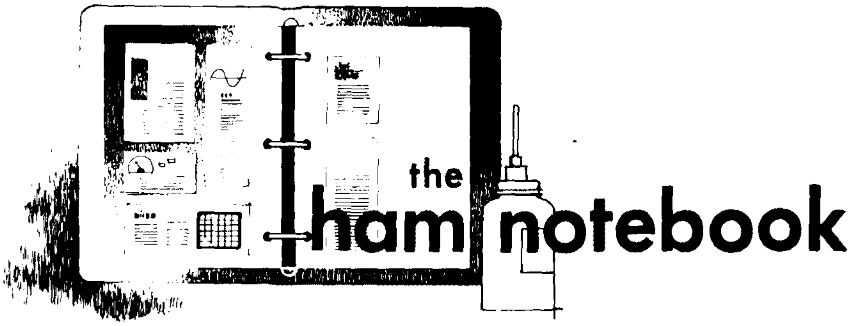
GMT	MST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CST
0000	5:00	30	30	15	10	15	10	10	20	6:00
0100	6:00	30	30	15	10	15	10	10	20	7:00
0200	7:00	30	30	15	10	15	10	10	20	8:00
0300	8:00	30	30	15	10	15	10	10	20	9:00
0400	9:00	30	30	20	10	15	10	15	20	10:00
0500	10:00	30	30	20	10	15	10	15	20	11:00
0600	11:00	30	*30*	20	15	15	15	15	30	12:00
0700	12:00	30	*30*	20	15	*20*	15	15	*30*	1:00
0800	1:00	*30*	*30*	20	20	*20*	15	20	*30*	2:00
0900	2:00	*30*	*40*	*20*	*20*	*20*	20	20	*30*	3:00
1000	3:00	*30*	*40*	*20*	*20*	*20*	20	*20*	*30*	4:00
1100	4:00	*30*	*40*	*20*	*20*	*20*	*20*	*20*	*30*	5:00
1200	5:00	*20*	*30*	*20*	*20*	*20*	*20*	*20*	*30*	6:00
1300	6:00	*20*	*30*	*15*	*20*	*20*	*20*	*20*	*40*	7:00
1400	7:00	*20*	*30*	*15*	20	*20*	*20*	*20*	*40*	8:00
1500	8:00	20	*30*	15	20	*20*	*20*	*20*	*40*	9:00
1600	9:00	20	20	15	20	*20*	15	*20*	*40*	10:00
1700	10:00	30	20	15	20	*20*	15	15	*30*	11:00
1800	11:00	30	20	10	15	15	15	15	*30*	12:00
1900	12:00	30	20	10	15	15	10	15	20	1:00
2000	1:00	30	20	10	15	15	10	15	20	2:00
2100	2:00	30	20	10	10	15	10	10	20	3:00
2200	3:00	40	30	15	10	15	10	10	20	4:00
2300	4:00	40	30	15	10	15	10	10	20	5:00

EASTERN USA

GMT	EST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	7:00	30	30	15	10	15	10	10	20
0100	8:00	30	30	15	10	15	10	10	20
0200	9:00	30	30	15	10	15	10	10	20
0300	10:00	30	30	15	10	15	10	15	20
0400	11:00	30	30	20	15	15	15	15	20
0500	12:00	*30*	30	20	15	15	15	15	30
0600	1:00	*30*	*30*	20	15	20	15	15	*30*
0700	2:00	*30*	*30*	20	15	*20*	20	*20*	*30*
0800	3:00	*30*	*40*	20	20	*20*	*20*	*20*	*30*
0900	4:00	*30*	*40*	20	*20*	*20*	*20*	*20*	*30*
1000	5:00	*30*	*40*	*20*	*20*	*20*	*20*	*20*	*30*
1100	6:00	*30*	*40*	*20*	*20*	*20*	*20*	*20*	*30*
1200	7:00	*20*	*30*	*15*	*20*	*20*	*20*	*20*	*40*
1300	8:00	*20*	*30*	*15*	*20*	*20*	*20*	*20*	*40*
1400	9:00	*20*	*30*	15	*20*	*20*	*20*	*20*	*40*
1500	10:00	20	20	15	20	*15*	*20*	*20*	*40**
1600	11:00	20	20	15	20	*15*	15	*15*	*40*
1700	12:00	30	20	10	15	*15*	15	15	*30*
1800	1:00	30	20	10	15	15	15	15	*30*
1900	2:00	*30*	20	10	15	15	10	15	*30*
2000	3:00	*30*	20	10	10	15	10	10	20
2100	4:00	30	20	10	10	15	10	10	20
2200	5:00	40	30	15	10	15	10	10	20
2300	6:00	40	30	15	10	15	10	10	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during 'normal' hours.

*Look at next **higher** band for possible openings.



KWM 380 external control circuit

This unit will control external antenna switches or solid-state (relay-actuated) amplifiers from the megahertz BCD output signals from the KWM 380. It was built in an evening with wire wrapping techniques using IC sockets, parts, and boards available at any Radio Shack or similar electronics outlet.

The KWM 380 (**fig. 1**) powers the circuit and provides BCD information for both the ones and ten digits. When operation is below 10 MHz, both of the "tens" BCD outputs, representing 10 and 20 MHz, are high.

The two 4028 IC's are BCD to decimal decoders. One handles the "ones" input information from the 380; the other handles the "tens" input information. The output from these IC's labeled U3 and U4 go to common QUAD AND gates U1 and U2 which are 74LS08's. When both inputs of any one of the AND GATES are high, the output goes high. This output goes through a 4.7k resistor to the base of a 2N2222 transistor. When the base goes high, the transistor acts as a switch and completes the circuit to ground for the small relay which in turn actuates an antenna switch (or amplifier bandswitch).

U1 and U2 pin 14 is at +5 volts, and pin 7 is at ground. U1a and U1b have one input each tied to the decimal 3 output of BCD decoder U4. This is because the KWM 380 causes both interface pins 12 and 13 to go high (10's and 20's MHz info) when the radio is tuned to any MHz band below 10.000. Thus U1a, which closes the 3.5 MHz relay, reads 33 MHz from the decoders, and U1b reads 37 MHz to close the 7 MHz relay. U3 and U4 pin 16 is at +5 volts,

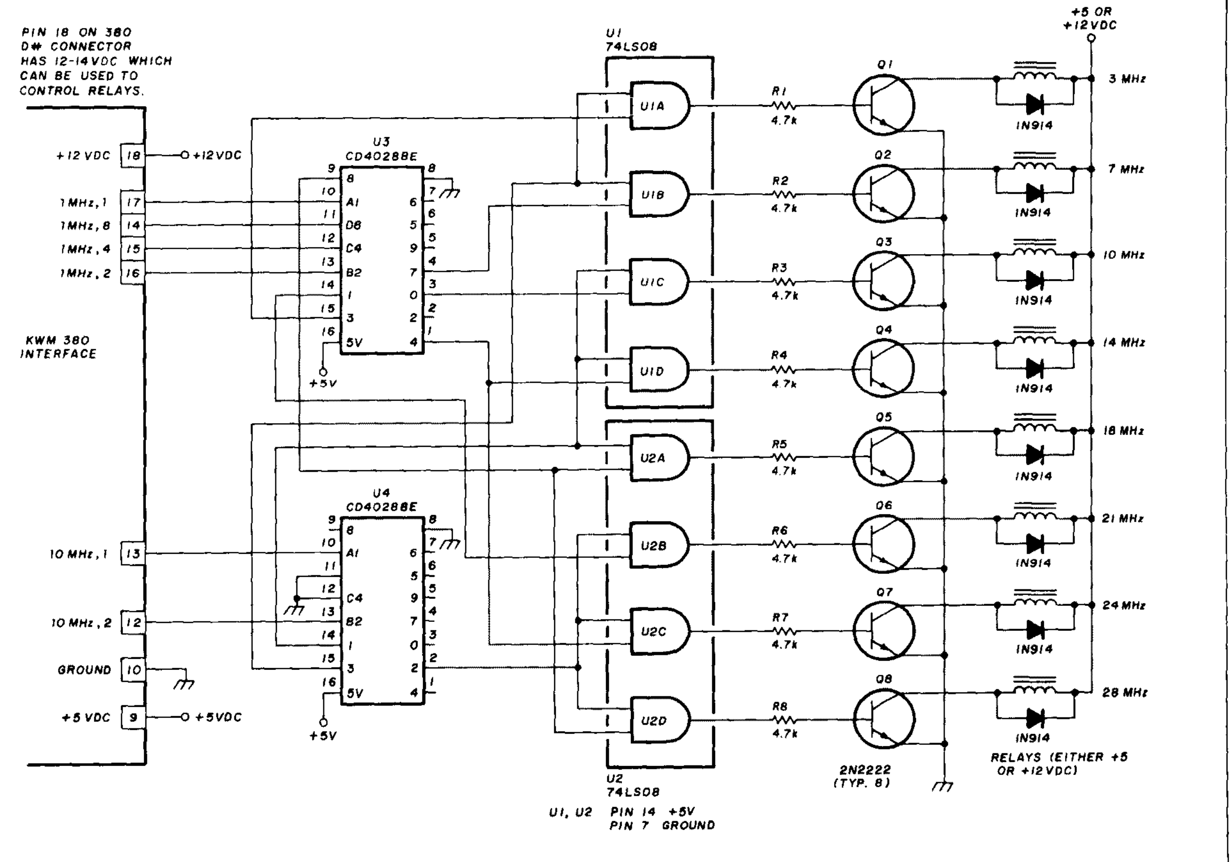


fig. 1. External control circuit schematic for the KWM-380.

and pin 8 is grounded. Unused BCD inputs U4 pins 11 and 12 must be grounded to prevent instability.

I recommend that the cable from the KWM 380 to the control unit be shielded to prevent EMI. The control circuit can be expanded to as many as 29 AND GATES if necessary to control the bandswitching systems required by some commercial solid-state amplifiers. Bypass capacitors of 0.1 and 0.01 μF should be added to the 5-volt and 12-volt lines to help reduce RF problems that might occur under high power conditions.

Bill Levy, WA2RUD

turns per inch from wire size

Here's a nice little way of computing the approximate number of turns per inch of enameled wire (of a tightly-wound single-layer coil) if you know the wire gauge.

$$TPI = \left(\left[\frac{G^3}{415} \right] + 5 \right)$$

where G is the wire gauge.

It's easy to put into a computer.

Dennis Mitchell, K8UR

regulator problem solved

For years I had been building 12 volt DC power supplies using the LM-340-K-12 and other three-legged devices without any apparent problems. But suddenly I started having problems, and blew out a dozen of the devices, which the manufacturer swears is well protected and easy to use.

I checked the input voltage to the regulator carefully; it was well below safe limits. The bypass capacitors were soldered close to the input and output leads, and I'd used a heatsink while soldering to the pins, but the things still blew out.

I wrote to several manufacturers and finally received an answer. Most of the circuits shown in the handbooks and brochures don't show a safety diode placed across the unit from output to input. It seems that when the voltage is removed from the input, the output capacitor discharges through the regulator and damages it. The solution, suggested in fig. 1, seems to solve the problem.

Ed Marriner, W6XM

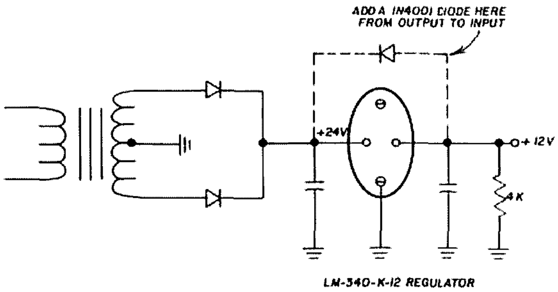
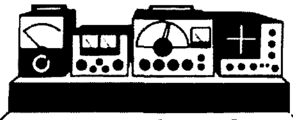


fig. 1. Including a diode across the regulator keeps it from "blowing."



product REVIEW

Fluke Model 8026 True RMS Multimeter

My first experience with Fluke equipment came while I was stationed at Letterkenny Army Depot in Pennsylvania. Letterkenny is the prime repair center for the Nike-Hercules and Hawk/Improved Hawk Missile Systems. Our repair and overhaul was a soup-to-nuts mission in that besides rebuilding the actual missiles themselves, we also overhauled all of the electronics systems.

As a member of the Production Engineering Department, I was directly involved in the establishment of the overhaul lines and was responsible for the selection and recommendation of test and production equipment. A number of overhaul steps required specific pieces of test equipment to ensure that all test procedures were done in accordance with manufacturers' specifications. This entailed establishing test and overhaul equipment lists that duplicated exactly what the systems' manufacturers used. This is where I first came into contact with Fluke equipment, which was specified because of its high degree of design accuracy and dependability in a production environment.

When the Fluke Model 8026B Multimeter came in to *ham radio* for review, I was quite interested to see how the unit compared to my memory of Fluke lab grade (read *very expensive*) test equipment. Needless to say, I wasn't disappointed. The Fluke Model 8026B is a handy, easy-to-use, digital multimeter. It is designed to test the following parameters:

DC voltage	100 μV - 1000 V
AC voltage	100 μV - 750 V
DC current	1 μA - 2000 mA
AC current	1 μA - 2000 mA
Resistance	0.1 ohm - 20 megohm
Diode test	
Conductance	0.1 ns - 200 ns 0.001 ms - 2 ms s = siemens = 1/ohm
Continuity	

It will also give true RMS AC measurement for signals up to 10 kHz. Each range has full auto-polarity operating, overrange indication and is protected from overloads. To ensure that measurements are accurate and noise free, the Model 8026B utilizes a dual slope integration measurement technique.

The unit measures 7.1 × 3.4 × 1.8 inches (180 × 86 × 45 mm) and weighs just 13 ounces (369 grams). With an alkaline 9 volt battery, the unit is rated at 200 hours of continuous operation. An optional AC power supply is available. The unit fits comfortably into your hand and the case is made of a high impact plastic.

operation

The unit is designed around a Fluke custom LSI chip. The chip contains a dual slope a/d converter and a driver for the LCD display. When an input signal is applied, it is routed through the range switch or to one of four signal conditioners as determined by the function switch setting. The conditioners are designed to scale and convert the input to an acceptable −0.2 to +0.2 VDC that is presented to the a/d converter.

Tuning for the Model 8026 is derived from a precision quartz crystal whose frequency is a multiple of 60 Hz. This allows the conditioned DC input to be measured over an integral number of power line cycles.

Of significant interest is the 103-page owner's manual. It is well written and full of information and goes far beyond owners' manuals found in many other units. Full documentation includes specifications, operating and maintenance instructions, a discussion of its theory of operation, a list of replaceable parts, accessory information, and a schematic.

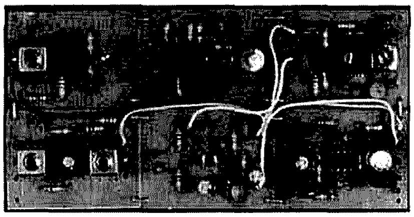
As you would expect from a company such as Fluke, calibration is a simple process and can be obtained from any Fluke Technical Service Center for a small fee. With all that Fluke has provided, it is hard to believe that I was reviewing a hand-held multimeter. The price of $219.99 may be a little expensive for the average ham. However, for the ham who wants only the very best or for the service technician who depends upon his equipment to perform without problems, I would highly recommend this or any other Fluke unit. For more information contact the John H. Fluke Mfg. Co., PO Box C9090, Everett, WA 98206. R.S.#302

N1ACH

NEW products

hobby kits

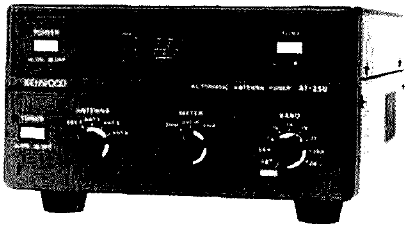
"Hobby Kits" are a new line of custom-designed electronic products from Morning Distributing. Designed with the ham or electronic hobbyist in mind, they allow construction of equipment ranging from a simple intercom or



phone amplifier to projects as complex as a state-of-the-art multi-band, multi-mode transceiver. Using this simple approach, beginners will be able to learn as they build. For engineers, it will allow them tremendous flexibility in equipment design. Modules are typically 1.5 to 2 inches long and 1.5 inches wide. All Hobby Kit modules other than the power supply are designed to work from 12-14 VDC. Boards are all predrilled and tinned. The kits come with all parts and instructions necessary. The photo above shows a monoband QRP transceiver built with six Hobby Kit modules for a total cost of about $35. A complete manual showing all diagrams and board layouts including suggested circuit systems design and hookups is available for $2.00 from Morning Distributing Company, P.O. Box 717, Hialeah, Florida 33011. RS#301

automatic antenna tuner

While the new AT-250 automatic antenna tuner from Trio-Kenwood has been specifically designed to match Kenwood's popular TS-



NEW products

430S high-frequency transceiver in size, color, and general appearance, it is functionally compatible with any HF transceiver of 200 watts PEP or less. Used with the TS-430S, its ABC (Automatic Band Change) system handles all switching from band to band; if the transceiver is other than the TS-430S, manual switching from band to band is required. The unit covers 160-10 meters, including the three WARC bands, has a front panel SWR/power meter, features four separate antenna terminals, and comes complete with a built-in AC power supply.

Additional information is available from Trio-Kenwood Communications, 1111 West Walnut Street, Compton, California 90220.

pocket oscilloscope

Calvert Instruments has developed a 4-ounce, pocket-size solid-state digital LED oscilloscope that provides the same functions as oscilloscopes more than twenty times its size, yet performs with only four solid-state controls. Focus and brightness on the 210-point, high-intensity illuminated screen are electronically self-controlled. The trace is always in sharp focus. Zero position and full-screen sweep are maintained automatically. The zero-reference, or cross-over line, is always centered. Full-

trace minimum on the screen is automatically maintained, and the automatic internal circuitry assures a properly positioned wave form.

The Pocket-O-Scope digital display visually reveals the value of the incoming signal during a precise real-time and/or degree envelope of the signal cycle. Every signal received is displayed as a series of lighted LED dots on the screen. These LED dots are for the "Digital Dot Envelope" and represent the digital value of

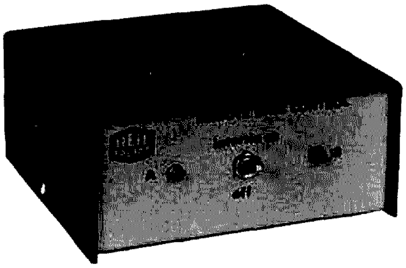
the signal rather than the analog value seen on conventional scopes. The Pocket-O-Scope's digital display provides a definite value for amplitude and its intra-relationship with real time/degrees. No judgment is required.

The Pocket-O-Scope, including carrying case, AC adapter, standard and high voltage probes and 200-page training manual, is being introduced for under $250. The scope only with standard probes will be available for $179.95.

For further information, contact Calvert Instruments, 19851 Ingersoll Drive, Cleveland, Ohio 44116. RS#303

remote base intertie

The RB-1 from Heil, Ltd., allows the easy interconnection of two transceivers for remote base operation. With the RB-1, a 220 MHz rig can be intertied with a 146 MHz rig so that the operator of a 220 MHz hand-held unit can control the 146 MHz "base" rig, which can be connected to large amplifiers, antenna systems, etc. This allows the HT to communicate over greater distances.



Very simple to connect and operate, the RB-1 has only two 8-pin microphone type connectors on the back panel to connect the squelch of rig A to the PTT Line of rig B, along with proper logic and audio coupling. With the front panel switch off, both rigs operate normally. The RB-1 can also be used as a repeater control.

The size of the unit is 4-1/4 × 4-1/8 × 1-7/8 inches; the price is $49.95.

For further information, contact Heil, Ltd., Marissa, Illinois 62257. RS#304

amplifier and power supply

The SCA-100 is a 100 watt, 406-512 MHz amplifier that can be used in either repeater or base station applications. With a unique high efficiency cooling design that combines a high volume forced air system and deep fin heatsink, failures due to transistor overheating are virtually eliminated.

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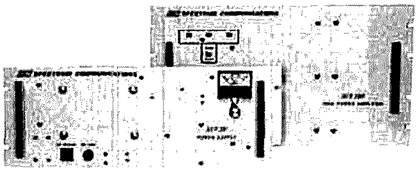
NEW products

The SCA-100 also features automatic high VSWR shutdown designed to self-test the antenna circuit in order to prevent damage to the amplifier should the antenna system fail. Other standard features include power supply failure



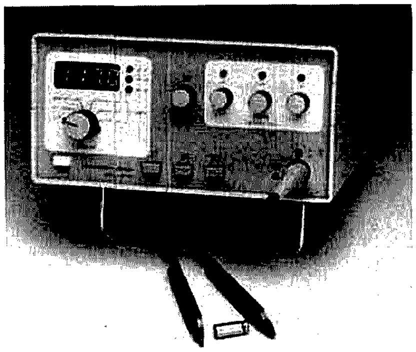
bypass to ensure continuous operation, over-temperature protection in case of cooling system failure and extremely tight RF shielding.

The SCP-30, the matching power supply for the SCA-100 amplifier, is designed to operate at 25 amps continuous, 13.8 VDC. Its features include ferro-resonant power transformer, over-current protection, optional power main failure switch to battery power and heavy-duty construction.

For further information, call Spectrum Communications Corp., 1055 West Germantown Pike, Norristown, Pennsylvania 19401-9616. RS#305

autoranging L/C meter

An instrument that allows accurate measurement and sorting of inductors and capacitors is available from Cambridge Technology. Although less than $1000, the Model 520 Com-



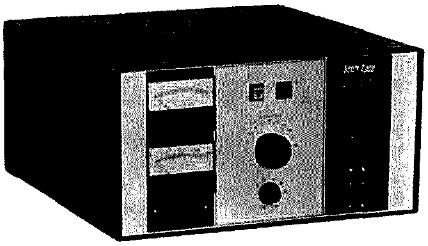
parator> <Bridge measures component dissipation and is accurate to 0.25 percent of reading. A built-in comparator allows simultaneous sorting for high and low tolerances, and dissipation limits. The limits are easily set with 10-turn controls and dual-color LEDs that indicate whether the component parameters fall within the ranges set.

The Model 520 may be easily calibrated by the user without any standard components, and has an internally adjustable, 0 to 10 VDC, capacitor bias voltage. Measurement frequencies of 120 Hz or 1 kHz are automatically selected. The instrument autoranges over unusually large measurement range from 199.9 pF/μH full scale to 1999 μF/H full scale. A range-hold feature permits fast, repetitive, true 4-terminal measurements. A full refund within thirty days is available if the instrument is found unsatisfactory for any reason. The Comparator> <Bridge sells for $785.00.

For more information, contact Cambridge Technology, Inc., 2464 Massachusetts Avenue, Cambridge, Massachusetts 02140. RS#306

new amplifier

Several new amplifiers are available from Henry Radio. The 2002A, a new version of the popular 2002 2-meter amplifier, uses the new Eimac 3CX800A7 power triode. The RF chassis



uses a 1/4-wavelength stripline design for a simple, straightforward, and reliable design. Like its predecessor, the 2002A offers 2000 watts input for SSB and 1000 watts input for CW. Because the tube provides more than 15 dB of gain, only about 25 watts drive is required for full output. The 2904A will be identical to the 2002A except that it is set up for the 430 to 450 MHz band, using a 1/2-wavelength stripline.

NEW products

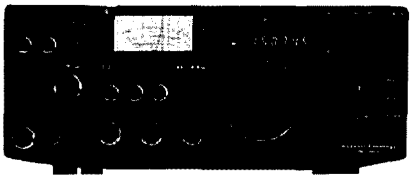
The 1002A, a 2-meter amplifier, follows the same design as the 2002A, but uses an 8874 tube for one-half the power specifications. The 1002A is rated at 600 watts PEP output and 300 watts continuous carrier output. It employs the same stripline design as the 2-02A. The 1004A, a half-power version of the 2004A, also uses the 8874. The 1004A will cover the 430 to 450 MHz band using 1/2- wavelength stripline design.

For further information, contact Henry Radio, 2050 South Bundy Drive, Los Angeles, California 90025. RS#307

HF transceiver

The new IC-745, a full feature HF transceiver and general coverage receiver, is now available from ICOM. Operational modes are SSB, CW, RTTY, AM (receive only) and FM (optional).

The unit offers the user the capability of a general coverage receiver, between 100 kHz and 30 MHz, and all of the HF ham bands from



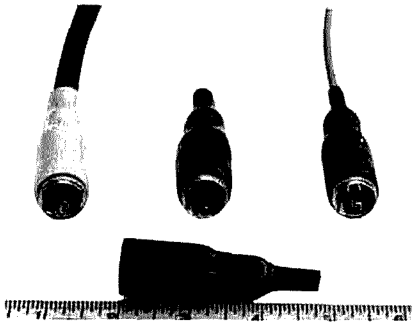
1.8 to 30.0 MHz including the new WARC bands at 10 MHz, 18 MHz, and 24 MHz. Ham band selection can be activated by simply touching the band button and rotating the main tuning knob. Other innovative standard features found in the IC-745 include 16 tunable memory channels, passband tuning, continuously adjustable AGC, 100 percent duty cycle-rated transmitter, and 12 volt DC operation.

A multi-purpose scanner allows the user to search the 16-memory channel frequencies or scan between two programmed frequencies. The 16 tunable memory channels have the capacity to store not only the desired frequency, but also the desired mode of operation. The frequency called up from memory can be changed by simply adjusting the frequency dial. Installation of the optional IC-PS35 internal power supply makes the IC-745 self-contained.

For further information, contact ICOM America, Inc., at 2112 116th Ave. N.E., Bellevue, Washington 98004. RS#308

new boots

Kilo-Tec announces a new custom weather boot for use with RG-8X, RG-59, RG-58, and RG-213 coaxial cables and PL-259/SO-239 combinations. Designed to keep connections

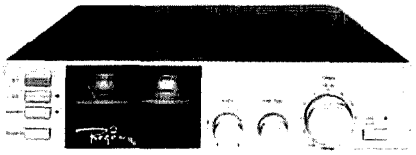


clean and dry and to keep moisture out of coax cables, the boots are manufactured with a flexible vinyl material that resists moisture and breakdown from the sun's rays. Three models are available: model KTB-58 for RG-58 cables; model KTB-8 for RG-8 cables; and model KTB-8X for RG-8X or RG-59. Custom weather boots can be made for other types of cables and connectors.

For further details, contact Kilo-Tec, P.O. Box 1001, Oak View, California 93022. RS#309

satellite TV receiver

The new Regency satellite receiver links your television set to its outdoor satellite antenna, or "dish." The Model SR 3000 receiver/downconverter combination is capable of receiving programming from a choice of satellites which include some 90 channels covering entertainment, sports, news, religious, and educational media. Its features include automatic Chaparral Polarotor control, detent tuning with AFC, pre-

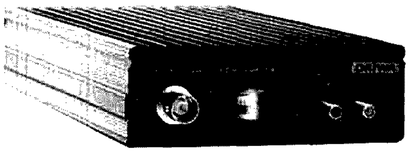


set and variable audio control, signal strength and center tuning meters, audio and video fine tuning controls, and a rugged weather sealed downconverter.

The suggested retail price on the Regency SR 3000 is $549.95. For further information, contact Regency Electronics, 7707 Records Street, Indianapolis, Indiana 46226. RS#310

two-meter linear amplifier

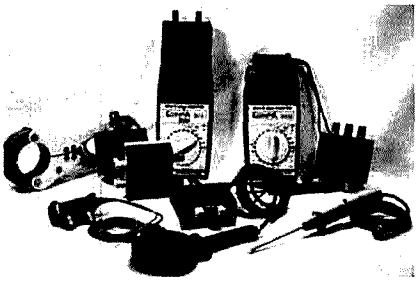
The LA2060, Daiwa's new medium power two meter linear amplifier, can take the output of a typical handheld transceiver and boost it to as much as 60 watts with any FM, SSB, or CW input from 0.5 to 3 watts acceptable. Equipped with RF activated transmit/receive switching (manual override option included), the unit features automatic protection circuitry and relative output metering. An input cable with BNC plugs is provided. The LA2060 requires 12 VDC at 12 amps maximum.



For details, contact MCM Communications, 858 E. Congress Park Drive, Centerville, Ohio 45459. RS#311

digital multi-multimeter™

The new AEMC digital multi-multimeter™ functions as a highly accurate voltage, current, and resistance tester designed for safe and precise testing of both sensitive electronic circuitry and high capacity power distribution networks. Its wide measuring capabilities include 24 ranges covering the following spans: 100 micro-volts DC to 1000 volts DC; 10 milli-volts AC to 1000 volts AC; 1 micro-amp DC to 10 amps DC; 10 milli-amps AC to 10 amps AC; and 0.1 ohms to 2 megohms. An extra large LDC digital display (0.7 inch) allows for easy reading. Two models of the multi-multimeter™ are available: Model 2010, the average sensing conversion type, and Model 2011, which is true RMS.



Twelve interchangeable modules are now available for the unit. By using the various plug-in modules in place of the standard plug-

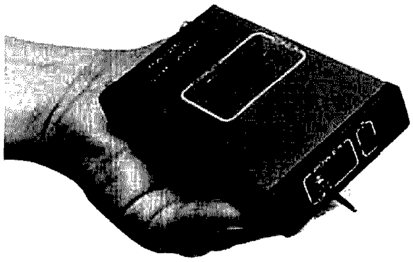
NEW products

in lead block, it is possible to measure capacitance, temperature, power for fiber optics, sound level, HVAC air flow, AC and DC current, frequency and light — as well as simulating RTDs, thermocouples, and process signals.

Further information is available from AEMC, 99 Chauncy Street, Boston, Massachusetts 02111. RS#312

compact amplifier

After nine years of manufacturing electronic inspection systems, Ham Industries, Inc., has expanded its product base to include several newly developed products designed for the Radio Amateur.



The first product to be released in the PA-25, a compact 25-watt amplifier for the 2-meter band that boosts the output power from a handheld transceiver by a factor of *six*. The device weighs only eight ounces and can be attached directly to a handheld unit or permanently mounted to the automobile dashboard. An adapter cord allows plugging directly into the car cigarette lighter; a separate power supply can also be used.

For more information, contact Ham Industries, Inc., 835 Highland Road, Macedonia, Ohio 44056. RS#313

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flea market

WANTED, MILITARY SURPLUS RADIOS. We need Collins 618T, ARC-72, ARC-94, ARC-102, RT-712/ARC-105, ARC-114, ARC-115, ARC-116, RT-823/ARC-131, or FM-622, RT-857/ARC-134 or Wilcox 807A, RT-743/ARC-51A, ARC-159, RT-1167 or RT-1188/ARC-164, ARC-186, RT-1022/ARN-84, ARN118, RT-859/APX-72, RT-868/APX-76, APN-153, APN-155, APN-171, 718F-1/2, RT-618/URC, AM-3007A/URT, Collins antenna couplers type 490T-1, 490T-2, 490T-9, CU-1658A/ARC, CU-1669/GRC, 490B-1, 690D-1, CU-1239/ARC-105. Top dollar paid or trade for new amateur gear. Write or phone Bill Slep 704-524-7519, Slep Electronics Co., Hwy. 441, Otto, N. C. 28763.

WANTED: Cash paid for used Speed Radar equipment. Write or call: Brian R. Esterman, P.O. Box 8141, Northfield, Illinois 60093. (312) 251-8901.

WANTED: Old RCA, Western Electric tubes. (713) 728-4343. Maury Corb, 11122 Atwell, Houston, Texas 77096.

PARABOLIC ANTENNA, spun aluminum, 6 ft. with mount $325.00. 408-730-2500. Norman, 2225 Sharon Rd., #224, Menlo Park, CA 94025. (415) 854-0266.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

WANTED: Boonton/HP type 250B RX meter also service manual for ex USN receiver type R1051B/URR. Write: Gill, 72 Elgin Street, Gordon, 2072, Australia.

RADIO WEST! High quality coverage communications receivers with Collins filters and other DX modifications! Catalog 50¢ (refundable). Radio West, 3417 Purer Rd., Dept. HR, Escondido, CA 92025. (619) 741-2891.

VERY In-ter-est-ing! Next 4 issues $2. Ham Trader "Yellow Sheets", POB356, Wheaton, IL 60189.

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Coming Events

ACTIVITIES
"Places to go..."

WISCONSIN: The West Allis Radio Amateur Club's 12th annual Midwinter Swapfest, Saturday, January 7, Waukesha County Expo Center Forum. Starting time 8 AM. Admission $2.00 advance; $3.00 at door. Reserved tables $3.00 until 11 AM or $4.00 at the door. Delicious food. For tickets or information SASE to WARAC, PO Box 1072, Milwaukee, WI 53201.

South Bend, INDIANA: Hamfest Swap & Shop, January 8, first Sunday after New Year's Day at Century Center downtown on U.S. 33 Oneway North between St. Joseph Bank Building and river. Industrial history Museum in same building. Carpeted half acre room. Tables $3 each. Four lane highways to door from all directions. Talk in freq: 52-52, 99-39, 93-33, 78-18, 69-09, 145.43, 145.29.

OPERATING EVENTS
"Things to do..."

DECEMBER 16 AND 17: The Triple States Radio Amateur Club will operate WD8DDL/8 from Bethlehem, West Virginia, 1400 to 0200 UTC daily. Frequencies: 7.275, 14.325, 21.425 and 28.550 MHz on SSB, 7.110, 14.075, 21.110 and 28.110 MHz on CW and 14.095 MHz on RTTY. A special holiday certificate will be sent to all those contacted who send a SASE to TSRAC, 26 Maple Lane, Bethlehem, Wheeling, WV 26003.

DECEMBER 17: The Burn family will operate B-N8EVE, U-KD8EO, R-N8EVF, N-KD8EV during Seasons Greetings from "The Little Town of Thompson, Ohio," from 1400-2000 UTC and December 18 from 1400-2000 UTC on 40 meters in General class band. You must work all four hams to receive your certificate. Please send SASE and QSL to 6215 Clay Street, Thompson, OH 44086 for 8½ × 11" certificate.